CHAOTIC COMMUNICATION SYSTEM AND METHOD
USING MODULATION OF NONREACTIVE CIRCUIT ELEMENTS

This application is a continuation-in-part of commonly-owned, copending U.S. application serial number 09/116,661, now abandoned entitled "Communications System Using Chaotic Synchronized Circuits," filed on July 17, 1998 and naming as inventors Daniel E. Hinton, Sr. and Aaron Budgor. That application is incorporated into the body of this application with minor changes in the numbering of elements in the figures.

TECHNICAL FIELD

This invention relates generally to information transmission techniques involving modulation and demodulation of a chaotic carrier signal. Many aspects of the invention involve transmitting information by modulating various characteristics of nonreactive circuit elements of a chaotic transmitter. The invention has broad application to communications systems, radar systems and other systems that transmit and receive information over wire, radio frequencies, light (including fiber optic) and acoustic channels.

BACKGROUND OF THE INVENTION

Techniques for modulating carrier signals in order to transmit information between two points are well known. In systems employing frequency modulation, for example, a carrier signal is modulated by changing the frequency of the signal in accordance with an information signal such as a human voice. Amplitude-modulated systems change the amplitude of a fixed-frequency signal in accordance with an information signal. Other modulation techniques have been developed over the years to optimize transmission characteristics, to optimize signal bandwidth, and to overcome noisy transmission environments.

So-called "chaotic" signals provide a particularly interesting, simple, and useful means of modulating information signals in a manner that can increase noise immunity and reduce the power levels needed to transmit information. As explained in the aforementioned application, which is bodily incorporated herein, these signals can be modulated in various ways to transmit information. The modulation

1 bandwidth available when using such techniques, however, has been determined to

2 be generally limited to 10 to 15% of the tank circuit frequency in the transmitting

3 circuit. This limitation is believed to be due to the fact that changing lump

4 parameters in the transmitter causes a certain amount of settling time before the

5 receiver can synchronize with the changed transmitter parameters.

6 The present inventors have discovered a technique for modulating the

7 transmitting signal in a manner that results in much faster signal stability, thus

8 reducing the amount of time required to synchronize the receiver and increasing the

9 modulation bandwidth dramatically. Other features and advantages provided by the

10 present invention will become apparent upon reading this specification in conjunction

11 with the figures.

12 The following description begins by reviewing the subject matter of the

13 aforementioned application as a departure point for explaining the principles of the

14 present invention. Circuits, principles and embodiments described in the

15 aforementioned application will be referred generally to as "first-generation," while

16 those newly presented in this application will be referred to generally as "second-

17 generation" or "improved." These labels are not intended in any way to be limiting.

18 Moreover, many of the second-generation circuits and principles can be used in

19 conjunction with first-generation circuits and vice versa.

20 **BRIEF DESCRIPTION OF THE DRAWINGS**

21 FIG. 1A is a diagram of a Chua circuit according to the prior art.

22 FIG. 1B is a communications system according to the prior art.

23 FIG. 1C is a diagram indicating the resistance – voltage characteristic of a

24 non-linear resistor used in a Chua circuit.

25 FIG. 1D is a diagram of the operating regimes of a Chua circuit mapped into

26 a lump parameter plane.

27 FIG. 2A is a schematic of a transmitter of a first-generation system wherein

28 a capacitor 237 is switched to modulate a chaotic signal.

29 FIG. 2B shows a conventional receiver that can be used with the transmitter

30 of FIG. 2A.

1   FIG. 3A shows another embodiment of a first-generation transmitter that

2   produces a vocabulary of chaotic signals.

3   FIG. 3B shows a receiver for use with the embodiments of FIG. 3A and FIG.

4   4A.

5   FIG. 4A shows another embodiment of a first-generation transmitter that

6   produces a vocabulary of chaotic signals in which all such signals can be mapped to

7   the same combinations of the lump parameter plane of FIG. 1D.

8   FIG. 4B shows another embodiment of a first-generation receiver for use with

9   the embodiments of FIG. 3A and FIG. 4A, this receiver being usable with a simple

10   counter circuit for determining a beat frequency.

11   FIG. 4C shows another receiver for use with the embodiment of FIG. 3A and

12   FIG. 4B using a synchronizing resistor 385.

13   FIG. 4D shows another receiver usable with a simple counter circuit for

14   determining a beat frequency using a synchronizing resistor formed from a

15   combination of resistors.

16   FIG. 4E shows another receiver similar to that of FIG. 4D, but which adds an

17   emitter follower 353 to isolate the oscillator portion 361 from point 287.

18   FIG. 4F shows a receiver including a simple counter circuit for determining

19   a beat frequency, wherein resistors provide a synchronizing element to lock the

20   incoming voltage of the communications channel and the receiver generated voltage.

21   In this embodiment, voltage follower 363 isolates the receiver generated signal from

22   the incoming signal and allows the receiver generated signal to feedback into the

23   oscillator portion of the Chua circuit to cause faster synchronization.

24   FIG. 5 shows a generalized communications system with a synchronizing

25   filter 550 according to a first-generation embodiment of the invention.

26   FIG. 6A shows a conventional transmitter with a Kennedy non-linear diode

27   that can be modulated according to the principles of a second-generation system.

28   FIG. 6B shows a conventional transmitter with a Caltech non-linear diode

29   that can be modulated according to the principles of a second-generation system.

30   FIG. 6C shows a transmitter with a novel non-linear diode that can be

1   modulated in accordance with the principles of a second-generation system.

2   FIG. 7A shows how a switch 735c can be used to modulate voltage-current

3   characteristics of a nonlinear diode in accordance with a second-generation

4   embodiment.

5   FIG. 7B shows generally how an information signal 736 can be used to make

6   and break a switch 735c (or other switch-like device) to modulate a negative

7   resistance in accordance with a second-generation embodiment.

8   FIG. 7C shows the effects of changing various resistive values in a Caltech

9   diode (FIG. 6B) and Kennedy diode (FIG. 6A) on slope.

10   FIG. 8 shows how the slope of the current-voltage characteristic curve for a

11   nonlinear element can be changed in order to change the rotation of a strange

12   attractor phase plane according to a second-generation embodiment. FIG. 9A shows

13   modulation limits for Ga and Gb according to a second-generation transmitter.

14   FIG. 9B maps current-voltage characteristic curves between a modulated

15   transmitter and a receiver.

16   FIG. 9C shows how positive breakpoints can be established for a nonlinear

17   diode according to various second-generation embodiments.

18   FIG. 10 shows a field effect transistor 1001 placed across diodes 652 and 655

19   to implement on-off keying according to a second-generation embodiment.

20   FIG. 11 shows a gain control amplifier 1146 inserted into a receiving circuit

21   that reduces or eliminates the need for an automatic gain control (AGC) amplifier on

22   the input at point 1191.

23   FIG. 12 shows a general case of a gain control amplifier 1202 embedded in

24   a receiving circuit.

25   FIG. 13 shows a transmitter including an operational amplifier 1308 that

26   isolates a filter 1309 from the chaotic subsystem 222 according to a second-

27   generation embodiment of the invention.

28   FIG. 14A shows low-pass filtering characteristics of a transmitter according

29   to a second-generation embodiment of the invention.

30   FIG. 14B shows bandpass filtering characteristics of a transmitter according

1   to a second-generation embodiment of the invention.

2   FIG. 15 shows a receiver including noise filters 1593, 1594 and 1597 to filter

3   out noise components introduced by the transmission channel according to a second-

4   generation embodiment of the invention.

5   FIG. 16 shows a receiver including noise filters 1666, 1694 and 1697 to filter

6   out noise components introduced by the transmission channel according to a second-

7   generation embodiment of the invention.

8   FIG. 17 shows a receiver including filters 1793, 1766, 1794, 1797 and 1721,

9   wherein filter 1721 works with automatic gain control amplifier 1746 to further

10   reduce noise generated in subsystem 1726.

11   FIG. 18A shows a cell phone system incorporating a baseband modem using

12   chaotic modulation principles according to a second-generation embodiment of the

13   invention.

14   FIG. 18B shows a cell phone system using chaotic modulation principles at

15   the intermediate frequency level according to a second-generation embodiment of the

16   invention.

17   FIG. 18C shows a cell phone system using chaotic modulation principles at

18   the radio frequency level according to a second-generation embodiment of the

19   invention.

20   FIG. 19A shows a second-generation system employing various principles

21   of the present invention, including a transmitter with a nonlinear element 1904

22   modulated with an information signal, and a receiver including a nonlinear element

23   1905 and a synchronizing resistor Rsync.

24   FIG. 19B shows a current/voltage curve 1910 for a nonlinear diode

25   superimposed over a load line 1911.

26   FIG. 19C shows two "single-scroll" attractors 1920 and 1930 orbiting around

27   equilibrium points at the intersection of a nonlinear diode current-voltage

28   characteristic curve and a load line, illustrating a DC analysis of a transmitter.

29   FIG. 19D shows how the current-voltage characteristic curve can be changed

30   at the transmitter to move the equilibrium points between three different positions

1   while slope Ga and the breakpoint positions are held constant.

2   FIG. 19E shows a strange attractor moving with the equilibrium point as a

3   nonlinear circuit element is modulated with an information signal, causing changes

4   in the slope of part of the characteristic curve.

5   FIG. 19F shows one technique for changing a nonlinear diode current-voltage

6   characteristic curve using an ideal switch SW1 and a resistor in series with the ideal

7   switch that is also in parallel with one of the nonlinear diode resistors.

8   FIG. 19G shows the result of modulation on the voltage across the nonlinear

9   diode.

10   FIG. 20A shows a voltage (V1) versus voltage (V2) versus time (T) plot of

11   a chaotic signal (double scroll strange attractor) without modulation.

12   FIG. 20B shows the plot of FIG. 20A when modulated with an information

13   signal.

14   FIG. 20C shows a voltage (V2) versus current (I3) versus time (T) plot of a

15   chaotic signal (double scroll strange attractor) without modulation.

16   FIG. 20D shows the plot of FIG. 20C when modulated with an information

17   signal.

18   FIG. 20E shows a voltage (V1) versus current (I3) versus time (T) plot of a

19   chaotic signal (single scroll attractor) without modulation.

20   FIG. 20F shows the plot of FIG. 20E when modulated with an information

21   signal (single scroll modulation).

22   FIG. 21A shows a nonlinear diode current-voltage characteristic curve where

23   resistor R1 is set to 1200 ohms (double scroll attractor).

24   FIG. 21B shows a nonlinear diode current-voltage characteristic curve where

25   resistor R1 is set to 1210 ohms (double scroll attractor).

26   FIG. 21C shows a nonlinear diode current-voltage characteristic curve where

27   resistor R1 is set to 1220 ohms (double scroll attractor).

28   FIG. 21D shows a voltage-current phase space map (V1 vs. V2 vs. I3)

29   corresponding to the nonlinear diode curve of FIG. 21A.

30   FIG. 21E shows a voltage-current phase space map (V1 vs. V2 vs. I3)

1   corresponding to the nonlinear diode curve of FIG. 21B. As compared to FIG. 21D,

2   the strange attractor on the left side is "squashed."

3   FIG. 21F shows a voltage-current phase space map (V1 vs. V2 vs. I3)

4   corresponding to the nonlinear diode curve of FIG. 21C. As compared to FIG. 21E,

5   the strange attractor on the left side is even more "squashed."

6   FIG. 21G shows a frequency plot corresponding to the nonlinear diode curve

7   of FIG. 21A.

8   FIG. 21H shows a frequency plot corresponding to the nonlinear diode curve

9   of FIG. 21B.

10   FIG. 21I shows a frequency plot corresponding to the nonlinear diode curve

11   of FIG. 21C.

12   FIG. 22A shows a voltage (V1) versus voltage (V2) versus time (T) plot of

13   a chaotic signal (single scroll strange attractor) without modulation.

14   FIG. 22B shows the plot of FIG. 22A when modulated with an information

15   signal.

16   FIG. 22C shows a voltage (V2) versus current (I3) versus time (T) plot of a

17   chaotic signal (single scroll strange attractor) without modulation.

18   FIG. 22D shows the plot of FIG. 22C when modulated with an information

19   signal.

20   FIG. 22E shows a voltage (V1) versus current (I3) versus time (T) plot of a

21   chaotic signal (single scroll strange attractor) without modulation.

22   FIG. 22F shows the plot of FIG. 22E when modulated with an information

23   signal.

24   FIG. 23A shows a nonlinear diode current-voltage characteristic curve where

25   resistor R1 is set to 1930 ohms (single scroll attractor).

26   FIG. 23B shows a nonlinear diode current-voltage characteristic curve where

27   resistor R1 is set to 1940 ohms (single scroll attractor).

28   FIG. 23C shows a nonlinear diode current-voltage characteristic curve where

29   resistor R1 is set to 1950 ohms (single scroll attractor).

30   FIG. 23D shows a voltage-current phase space map (V1 vs. V2 vs. I3)

1   corresponding to the nonlinear diode curve of FIG. 23A.

2   FIG. 23E shows a voltage-current phase space map (V1 vs. V2 vs. I3)

3   corresponding to the nonlinear diode curve of FIG. 23B. As compared to FIG. 23D,

4   the strange attractor is elongated.

5   FIG. 23F shows a voltage-current phase space map (V1 vs. V2 vs. I3)

6   corresponding to the nonlinear diode curve of FIG. 23C. As compared to FIG. 23E,

7   the strange attractor is even more elongated.

8   FIG. 23G shows a frequency plot corresponding to the nonlinear diode curve

9   of FIG. 23A.

10   FIG. 23H shows a frequency plot corresponding to the nonlinear diode curve

11   of FIG. 23B.

12   FIG. 23I shows a frequency plot corresponding to the nonlinear diode curve

13   of FIG. 23C

14   FIG. 24 shows a dual-transmitter configuration (1200, 1205) according to a

15   second-generation embodiment of the invention.

16   FIG. 25 shows a dual-receiver configuration (601, 1370) according to a

17   second-generation embodiment of the invention.

18   FIG. 26 shows a subtraction circuit for detecting a voltage difference across

19   various points (e.g., point 1450) in FIG. 25.

20   FIG. 27 shows an absolute value circuit that can be used in conjunction with

21   a detector function.

22   FIG. 28 shows a dual receiver synchronization detector circuit according to

23   a second-generation embodiment of the invention.

24   FIG. 29 shows a dual receiver synchronization detector circuit in which

25   signals are subtracted, absolute valued and then subtracted.

26   FIG. 30 shows a detector circuit for detecting voltage changes between points

27   287 and 1415 or points 1440 and 1470 in FIG. 25.

28   FIG. 31 shows a baseband transmitter interface circuit for interfacing a

29   modulated chaotic transmitter to a communication system.

30   FIG. 32 shows a baseband receiver interface circuit for interfacing a receiver

1   to a communication system.

2   FIG. 33 shows a system in which a chaotic transmitter and receiver are

3   interfaced to an infrared amplitude modulated subsystem 3090.

4   FIG. 34 shows how a chaotic transmission and reception system can be

5   interfaced to a radio transmitter/receiver pair 3100 and 3110.

6   FIG. 35 shows a balanced cable driver circuit 2455 that can be used to pass

7   a chaotic signal over a twisted pair or coaxial cable system.

8   FIG. 36 shows a balanced cable receiver circuit 2880 that can be used to

9   interface a chaotic receiver 2860 to a twisted pair or coaxial cable system.

10   FIG. 37A shows curves representing chaotic operating regions for different

11   values of a synchronizing resistor 660 for a Caltech diode implementation (FIG. 6B).

12   FIG. 37B shows curves representing chaotic operating regions for different

13   values of a synchronizing resistor 608 for a Kennedy diode implementation (FIG.

14   6A).

15   FIG. 37C shows what happens when the transmitter capacitor 215 (FIG. 6B)

16   is varied and the receiver capacitors 355 and 1490 (FIG. 25) are set to fixed values

17   with the nonlinear diode characteristic curve set at a fixed value.

18   FIG. 38 shows a technique for doubling a signal rate using four unmodulated

19   oscillator/transmitters and corresponding receivers.

20   FIG. 39 shows a technique for increasing the digital signal transmission rate

21   using multiple chaotic transmitters and matched receivers.

22   FIG. 40 shows how a nonlinear circuit can be replaced with two functions that

23   represent only the Gb slopes, referred to as a "Gb-only" transmitter or receiver.

24   FIG. 41 shows a detector design in which a nonlinear diode is replaced with

25   +/- Gb slope detectors (5350 and 5340).

26   FIG. 42 shows a dual receiver design using sample-and-hold circuits with

27   outputs 5380, 5360, and 5370.

28   FIG. 43 shows a transmitter that modulates only the slope Gb.

29   FIG. 44 shows a dual-transmitter system that modulates only the slope Gb.

30   FIG. 45A shows a dual receiver design wherein a nonlinear diode is replaced

1   with a +/- Gb detector and voltage Vb in a negative resistor circuit.

2   FIG. 45B shows a dual receiver design that is a variation on that of FIG. 45A.

3   FIG. 46 shows a current-voltage characteristic curve for certain embodiments

4   of the invention that modulate and detect a positive slope.

5   FIG. 47 shows a receiver in which a nonlinear diode is replaced with a Gb+

6   detector.

7   FIG. 48 shows a Gb+ dual receiver design including a sample and hold

8   circuit.

9   FIG. 49 shows a Gb+ only transmitter using Gb+ slope modulation and

10   voltage modulation.

11   FIG. 50 shows a digital to analog Gb+ only transmitter.

12   FIG. 51 shows a current-voltage characteristic curve for certain embodiments

13   of the invention.

14   FIG. 52 shows a current-voltage characteristic curve for a positive Gb voltage

15   current M-ary modulation system.

16   **I. <u>FIRST-GENERATION EMBODIMENTS AND TECHNIQUES</u>**

17   Referring to FIG. 1A, a circuit 1 known as a "Chua" circuit oscillates

18   chaotically. The term "chaos" applies to dynamic systems that follow simple

19   dynamical rules, but whose state function trajectory is so sensitive to the system's

20   initial conditions that its state after an arbitrary time-period cannot, in practical terms,

21   be predicted. That is, its state could be predicted if it were possible to model the

22   system with an arbitrary degree of precision.

23   Chaotic systems evolve deterministically, and their chaotic state paths are

24   cyclic, but very complex and with extremely long cycle-lengths. In real systems,

25   however, with extremely long cycle periods, it may be of little practical significance

26   that their behavior is cyclical because the physical systems that generate the behavior

27   may not be sufficiently stable for the system to ever return to the same dynamical

28   system in its same initial state. For example, the component values of an electrical

29   circuit may not remain precisely constant for 600 years.

30   The Chua circuit is a simple electrical circuit that exhibits chaotic behavior.

1   It has been studied extensively and used to demonstrate many of the chaotic patterns

2   observed in many physical systems. Referring now also to Fig 1A, the basic Chua

3   circuit includes a non-linear resistance element 10, characterized by a non-linear

4   voltage-current characteristic curve. In a typical configuration, the curve is piece-

5   wise linear with symmetrical slope discontinuities around the zero-axis. That is: I_R

6   $= G_a V_R + (\frac{1}{2})(G_a - G_b)\{|v_R + B_p| - |v_R - B_p|\}$ where G_a and G_b are the slopes of

7   respective linear portions of the piecewise-linear current/voltage curve characterizing

8   the non-linear resistor and B_p is the absolute value of the two voltage points at which

9   the discontinuities in the current/voltage curve lie as shown in FIG. 1C. The circuit

10 has a circuit-driving subsystem 2 (e.g., an L-C tank circuit), and a response

11 subsystem 3, which includes for example a capacitance C1 and non-linear resistor 10,

12 wherein the two systems are interconnected through a resistor 25.

13   Referring to FIG. 1D, a given choice of values of the physical characteristics

14 of the components of the Chua circuit each correspond to a unique operating regime,

15 some values of which may coincide with a chaotic behavior of the Chua circuit. The

16 operating regime may be mapped onto a coordinate system whose axes are the lump

17 parameters, $\alpha = C_2/C_1$ and $\beta = R^2 C_2/L$. By choosing values of R (25), L (30), C_1 (15)and

18 C_2 (20) so that α and β lie in, for example, a double scroll region 60, a Chua circuit

19 can be made that will oscillate chaotically or quasi-periodically. Any point on the

20 plot corresponds to a different operating behavior and a selected point does not

21 exhaustively define a particular path of state trajectories. A selected point on the

22 curves can correspond to radically different behaviors depending on the initial

23 conditions.

24   Given a specified physical configuration and a specified initial state specified

25 by V_1, V_2, and I_L, the voltages across C_1 (15) and C_2 (20) and the current through L

26 (30), the evolution of the Chua circuit's state is deterministic, but chaotic. That is,

27 any Chua circuit with the same physical parameters and initial conditions will follow

28 the same course of states over time and this course will repeat itself over a very long

29 interval (perhaps many years). However, to an observer, the value of (for example)

30 voltage V_1 over a period of time shorter than this long interval looks like noise. Also,

1   initial states that differ only slightly can follow widely different state paths. In

2   addition, its power spectral density function is spread over a wide range of

3   frequencies, with a peak at the frequency of the fundamental of the L-C tank circuit

4   formed by L and C_2. However, compared to oscillators, such as used to generate

5   carriers for radio transmission, the peak is not pronounced; that is, it is very short and

6   wide.

7   The Chua circuit, aside from being a classic device for demonstrating,

8   studying, and modeling chaotic real-world systems, has also been proposed as a basis

9   for chaotic signal transmission. Generally a transmitting nonlinear dynamic circuit

10   produces a chaotic signal that can be used to induce a receiving chaotic system to

11   synchronize with it. The parameter of the transmitting chaotic circuit can be

12   modulated or perturbed responsively to an information signal. The parameter can be

13   a scalar, such as a voltage, tapped from the transmitting circuit and used as a signal.

14   The signal is applied to the receiving system, causing the receiving system to

15   synchronize with the transmitted signal. The chaotic signal from the synchronized

16   receiving circuit can be used with the modulated transmitted signal to recover the

17   information signal according to various prior art schemes. The chaotic signals that

18   can be derived from an oscillating Chua circuit are similar to spread-spectrum signals

19   including a range of frequencies. Chua circuits have been made to generate

20   communications signals in frequency bands ranging from audio to radio frequency

21   and in various media.

22   Various modulation schemes have been proposed. For example, a simple

23   signal summing system adds the information signal to the chaotic scalar. A more

24   complex correlation system uses a signal divider and multiplier at the transmitter and

25   receiver, respectively. In FIG. 1B, a prior art system uses a Chua circuit to transmit

26   signals and receive signals. The system has a transmitting Chua circuit 100 and an

27   identical (in terms of its chaotic oscillating properties) receiving Chua circuit 101.

28   The transmitting Chua circuit 100 oscillates in a chaotic or semiperiodic regime.

29   Generally, the two chaotic circuits 100 and 101 can be synchronized by

30   driving a portion of the receiving chaotic oscillator 101 with a driving function

1   tapped from the transmitting chaotic oscillator 100. L-C tank circuit 105 of the

2   transmitting Chua circuit 100 is linked through a resistor 81 to the capacitor/non-

3   linear resistor portion 106. The latter portion causes the oscillations of the L-C tank

4   circuit to become chaotic for certain values of the inductor 74, capacitors 71 and 73,

5   and resistor 81 as discussed above with reference to Fig 1D. The chaotic portion 108

6   of the identical receiving circuit 101, also a capacitor/non-linear resistor circuit,

7   reproduces the driving signal. That is, the transmitting 100 and receiving 101 circuits

8   follow precisely the same chaotic course of states (assuming no modulation is taking

9   place in the transmitting circuit 100).

10   It is known that the transmitting 100 and receiving 101 circuits will remain

11   synchronized even when a substantial amount of noise and/or information is injected

12   into the driving signal. Thus, in the prior art embodiment of FIG. 1B, a signal current

13   $I_i(t)$ is injected by a driver 76 that converts a signal voltage through an invertable

14   coding function $c(v_s(t))$. The decoded signal at the receiver is then obtained from the

15   received current signal $I_d(t)$ by applying the inverse coding operation to the received

16   current signal $I_d(t)$ to obtain a voltage signal containing the information signal.

17   Note that the term, "synchronous," in this context, characterizes the

18   convergence of two state variables toward identical or linearly related, but

19   continuously changing, sets of values. That is, a change in one variable corresponds

20   to a change in a synchronized variable that is linearly related to the change in the one

21   variable. Thus, plotting one variable against the synchronized variable over time, the

22   result, theoretically, is a straight line. Synchronization of non-linear systems, and the

23   mathematical modeling of such systems, is described in some detail in U.S. Patents

24   5,245,660; 5,473,694; 5,402,334; 5,379,346; 5,655,022; 5,432,697; and 5,291,555,

25   the entirety of each of which is incorporated by reference herein.

26   Prior art systems have been discussed widely, but few practical working

27   designs are known. The problems with practical synchronization systems are

28   summarized in the introduction of U.S. Patent No. 5,680,462. Synchronization

29   systems are inherently noisy and error prone due, at least in part, to the time it takes

30   for synchronization to occur in a noisy channel and because noise induces state

1   transitions in the receiver since it causes a breakpoint to be crossed. For example,

2   when a transmitting circuit is perturbed to encode a piece of information (a bit), it

3   takes a finite amount of time for the receiving circuit to begin to follow the trajectory

4   of the transmitted signal. Also, according to the prior art, modulation cannot span

5   too great a range. Otherwise, a tightly locked synchronization, which is, according

6   to the prior art, essential, cannot be maintained. In addition, the practical problems

7   attending achievement of high data throughput, the providing of reliable locking

8   performance, and various purely practical design considerations have not received

9   a great deal of attention. These prior art problems are addressed by the present

10   invention in both the first-generation and second-generation embodiments.

11   According to one aspect of a first-generation system, the invention provides

12   a spread-spectrum-like communications system that transmits information in a

13   chaotic signal. Other aspects of the invention include:

14   (a) a method for modulating a chaotic process to generate a signal to encode

15   information in the signal;

16   (b) a method for modulating a circuit that generates a chaotic signal in a

17   stable manner.

18   (c) transmitting and receiving chaotic circuits that are characterized by rapid

19   synchronization;

20   (d) a mechanism for imprinting and extracting information from two chaotic

21   devices synchronized by a chaotic signal in which the information is embedded such

22   that the chaotic signal can serve as the information carrier signal over a

23   communication channel;

24   (e) a communications system that permits the modulation of a chaotic process

25   so as to encode multiple independent streams of data on a same chaotic carrier signal,

26   in effect, implementing an N-word vocabulary, where N corresponds to a number of

27   stable chaotic oscillation states that are induced in a transmitter by modification of

28   a property of at least one of a resistance, a capacitance, and an inductance to tune an

29   oscillating circuit of the transmitter.

30   Briefly, an embodiment of the invention employs a transmitting oscillating

1   circuit capable of chaotic or quasiperiodic oscillation to generate a (chaotic or

2   quasiperiodic) carrier, preferably a voltage tapped through a voltage follower. A

3   property of the transmitting oscillating circuit, in an embodiment, an auxiliary

4   capacitance, is switched on and off to vary the capacitance of an L-C tank portion of

5   a Chua oscillator. The switching is controlled by an information signal to generate

6   a modulated chaotic signal.

7   Switching is performed with an optical isolator that requires zero output

8   biasing and introduces essentially no capacitance into the circuit. This prevents any

9   effect on the chaotic or quasi-periodic operating regime of the circuit. An

10   autonomous portion of a receiving oscillating circuit, substantially identical in terms

11   of its oscillating properties, is driven by the modulated carrier. This establishes a

12   synchronized chaotic or quasiperiodic oscillation in the receiving circuit. A

13   comparator is used to output the difference between the driving modulated carrier

14   and a synchronized signal tapped from the receiving oscillator at a point

15   corresponding to the transmitting circuit tap used to generate the modulated carrier.

16   This output provides the recovered information signal.

17   In one embodiment, various elements of the transmitting chaotic circuit are

18   switchably varied to maintain a constant operating regime so that strange attractors,

19   with frequencies covering a wide selectable range, are generated. This is used to

20   form a vocabulary of strange attractors. The frequency can be determined by the

21   receiver in a very simple way by counting pulses formed from a difference between

22   the base signal in the receiving circuit and the received signal. The difference in the

23   frequencies of the signal being transmitted and the base signal generated by the

24   receiver indicates the "word" transmitted. In this way, if the vocabulary consists of

25   N distinguishable oscillating frequencies, then $\log_2(N)$ bits can be transmitted with

26   each modulating cycle.

27   According to one embodiment, the invention provides a communications

28   device with a transmitting chaotic circuit. The transmitter has at least one circuit

29   element, the value of which affects a chaotic electrical property of the chaotic circuit.

30   That is, a change in the magnitude of the circuit element changes the oscillating

1 behavior of the chaotic transmitting circuit. The circuit element has multiple

2 component elements, at least one of which is isolated from the chaotic circuit by a

3 switch. The configuration is such that when the switch is switched to a first state, the

4 magnitude has a first value and when the switch is switched to a second state, the

5 magnitude of the component has a second value. This causes the transmitter to

6 oscillate over multiple oscillating regimes each corresponding to one of the values.

7 The chaotic property can be applied to a communications channel to be picked up by

8 a receiver.

9   Switching the circuit element allows, essentially, a chaotic signal to be

10 modulated. That is, a chaotic signal is tapped from the transmitter (in FIG. 1B, for

11 example, the voltage at the junction of resistor 81 and capacitor 71), applied to a

12 communications channel, and picked up by a receiver. The switch is controllable

13 responsively to an information signal, whereby the chaotic carrier signal is modulated

14 by the information signal. This information signal can be detected by applying the

15 chaotic signal from the channel to a receiving chaotic circuit that synchronizes with

16 the chaotic signal corresponding to one of the chaotic oscillating regimes of the

17 transmitter, but not with another and detecting the alternations between

18 synchronization and desynchronization.

19   According to another embodiment, the invention provides a communications

20 device with a transmitting chaotic circuit configurable responsively to an information

21 signal. The configurations are such that the transmitting chaotic circuit produces at

22 least three different chaotic signals, each characterized by a different

23 trajectory-versus-time characteristic. The device includes a receiver with an

24 oscillating subportion to which the different chaotic signals can be applied to drive

25 the oscillating subportion. The receiver has a beat detector connected to the

26 oscillating subportion to detect a difference between a fundamental frequency of the

27 oscillating subportion and a current chaotic signal. This allows the information

28 signal to be detected by the detection of beats.

29   According to still another embodiment, the invention provides a

30 communications receiver with a chaotic oscillator that includes an oscillator portion

1   and a chaotic portion. The chaotic portion has a non-linear resistance element that

2   forms a chaotic oscillator with the oscillator portion when the chaotic portion and the

3   oscillator portions are coupled to pass a current signal therebetween. The oscillator

4   portion is signally coupled to a communications medium carrying a modulated

5   chaotic signal. The chaotic portion is also signally coupled directly to the

6   communications medium such that a voltage of the communications medium is

7   directly applied to the chaotic portion through a circuit path parallel to a coupling

8   allowing the current signal to pass between the oscillator portion and the chaotic

9   portion. Thus, both the chaotic and oscillating portions of the receiver are driven by

10   the incoming chaotic signal from the communications channel.

11   According to still another embodiment, the invention provides a

12   communications device with a chaotic oscillator connectable to a communications

13   channel. The chaotic oscillator has a tank circuit with at least two capacitors and an

14   inductor. The first of the capacitors is connected to an inductor and a second is

15   selectively connectable to the inductor to combine respective capacitances of the

16   capacitors through a switch. In other words, the capacitors combine their capacities

17   responsively to the switch. The switch has an input for accepting an information

18   signal. The information signal controls the switch so that the chaotic oscillator is

19   selectively alternated between at least two oscillating regimes. The result is that a

20   chaotic transmitter is modulated in accordance with the information signal to

21   generate a chaotic signal which, at each instant, oscillates according to a selected one

22   of the oscillating regimes. A receiver signally coupled to the communications

23   channel has a receiving chaotic oscillator portion for each of the oscillating regimes,

24   each portion being configured to synchronize with a respective one of the at least two

25   chaotic signals. By detecting which portion is in synchrony with the incoming

26   signal, the information signal can be detected.

27   According to still another embodiment, the invention provides a

28   communications system with transmitting and receiving Chua circuits. At least one

29   component of the transmitting Chua circuit includes at least two subcomponents, at

30   least one of which is selectively isolated from the transmitting Chua circuit by a

1 switch. This is done such that a current oscillating regime of the transmitting Chua

2 circuit is selectively alternated between at least two respective oscillating regimes.

3 The switch is switchable responsively to an information signal. The values of the

4 subcomponents together with a configuration of the switch are such that one of the

5 oscillating regimes is substantially the same as an oscillating regime of the receiving

6 Chua circuit. The result of the latter is that the receiving Chua circuit is

7 synchronizable with the transmitting Chua circuit when the current oscillating regime

8 is the same oscillating regime as the receiver's. A detector is connected to detect

9 when the receiving Chua circuit is in synchrony with a chaotic signal generated by

10 the transmitting Chua circuit. This allows the information signal to be recovered

11 from the chaotic signal (see, e.g., FIG. 4C).

12 According to still another embodiment, the invention provides a

13 communications receiver with a chaotic oscillator that has an oscillator portion and

14 a chaotic portion. The chaotic portion has a non-linear resistance element that forms

15 a chaotic oscillator with the oscillator portion when the chaotic portion and the

16 oscillator portions are coupled to pass a current signal therebetween. The oscillator

17 portion is signally coupled to a communications medium carrying a modulated

18 chaotic signal. The chaotic portion is also signally coupled directly to the

19 communications medium such that a voltage of the communications medium is

20 directly applied to the chaotic portion through a circuit path parallel to a coupling

21 allowing the current signal to pass between the oscillator portion and the chaotic

22 portion. Thus, both the chaotic and oscillating portions of the receiver are driven by

23 the incoming chaotic signal from the communications channel. In this embodiment,

24 the coupling resistance is a series of three resistors that provide the coupling from the

25 voltage of the communications medium to the chaotic portion of the circuit and

26 provide a voltage divider network for a comparator detector to detect voltage

27 differences between the voltage of the communications medium and the chaotic

28 voltage generated by the receiver chaotic portion of the system (see, e.g., FIG. 4D).

29 According to still another embodiment, the invention provides a

30 communications receiver system as described in the previous paragraph except the

1 receiver is divided into an oscillator portion and a chaotic portion which are
2 separately driven through emitter followers from the voltage of the communications
3 medium. This allows the voltage of the communications medium to drive the
4 oscillator portion and the chaotic portion of the receiver without direct feedback
5 between the two through the synchronizing resistor. This prevents spontaneous
6 chaotic oscillation in the receiver due to a feedback path from the chaotic portion of
7 the Chua circuit to the oscillator portion of the Chua circuit. The circuit still
8 synchronizes since the voltage of the communications medium is coupled to both the
9 oscillator portion of the receiver and the chaotic portion of the Chua circuit. As a
10 result, the receiver responds only when there is a voltage on the communications
11 medium to stimulate the system (see, e.g., FIG. 4E).

12 According to still another embodiment, the invention provides a
13 communications receiver system divided into an oscillator portion and a chaotic
14 portion. The oscillator portion is driven by the voltage of the communications
15 medium. The voltage of the chaotic portion is fed back to the oscillator portion
16 through an emitter follower and the synchronizing resistor in a phase locking type
17 arrangement. This allows the voltage of the communications medium to drive the
18 oscillator portion and the voltage of the chaotic portion of the receiver to directly
19 feed back through the synchronizing resistor to quickly synchronize the
20 communications system in the presence of a voltage on the communications channel.
21 This arrangement rapidly synchronizes the communications system. The circuit still
22 synchronizes since the voltage of the communications medium is coupled to the
23 oscillator portion of the receiver and the chaotic portion of the Chua circuit feeds
24 back a portion of the receiver voltage. As a result, the receiver responds only when
25 there is a voltage on the communications medium to stimulate the system (see, e.g.,
26 FIG. 4F).

27 According to still another embodiment, the invention provides a
28 communications receiver system divided into an oscillator portion and a chaotic
29 portion except the nonlinear diode portion has only a Gb component. This allows the
30 oscillator portion of the circuit (e.g., FIG. 41 elements 5340 and 5350) to be driven

1 directly in accordance with the previous embodiments discussed above. By removing

2 the discontinuity caused by diodes 652 and 655 (FIG. 6B), noise in the channel

3 cannot cause the receiver to change scrolls. The circuit still synchronizes since the

4 voltage of the communications medium is coupled to both the oscillator portion of

5 the receiver and the chaotic portion of the Chua circuit. As a result, the receiver

6 responds only when there is a voltage on the communications medium to stimulate

7 the system (see, e.g., FIG. 41). The noise performance has been shown to be

8 approximately an 8-10 dB improvement over a receiver with the diodes in the circuit.

9 **DETAILED DESCRIPTION (FIRST-GENERATION EMBODIMENTS)**

10 Referring again to FIG. 1D, discussed above, the various chaotic oscillating

11 regimes of the Chua circuit are mapped onto an α/β parameter plane. Circuits falling

12 in a double scroll region 60 are characterized by oscillation about two strange

13 attractor equilibrium points ("double scroll"). Circuits falling in a spiral set of

14 oscillating regimes 61 exhibit oscillation about only one strange attractor equilibrium

15 point ("single scroll").

16 The oscillations of a circuit operating in a single-scroll attractor mode can be

17 seen graphically in FIG 22A, which plots two voltages V1 and V2 as a function of

18 time for a transmitter without modulation. The two voltages are measured at points

19 242 and 282, respectively, as shown in one embodiment of FIG. 2A. (In this

20 · example, V2 is measured at the L-C tank circuit and V1 is measured at the nonlinear

21 resistance element). While the circuit operates in a non-modulated state, the values

22 of V1 and V2 vary chaotically but generally swirl about an equilibrium point.

23 The oscillations of a circuit operating in a double-scroll attractor mode can

24 be seen graphically in FIG. 20A, which plots the same two voltages V1 and V2 as

25 a function of time for a circuit that is not modulated. While the circuit operates in

26 a non-modulated state, the values of V1 and V2 vary chaotically but generally swirl

27 about two distinct equilibrium points.

28 Varying C_1 causes the α/β combination to shift as indicated by arrow 130 in

29 FIG. 1D. Varying C_2 causes the α/β combination to shift as indicated by arrow 140.

30 Varying R_2 or L causes the α/β combination to shift as indicated by arrow 150. As

1   can be seen from the diagram (FIG. 1D), the capacitance C_2 of capacitor 20 can be

2   varied over a wide range while still maintaining operation of the circuit in the double

3   scroll oscillating regime 110. A much smaller range of values of capacitance of

4   capacitor 15 (C_1) coincides with operation in the double scroll regime 60. The wide

5   range of capacitances for C_2 (capacitor 20) that coincide with operation in the double

6   scroll region 60 is exploited in a first embodiment of the invention discussed

7   immediately below.

8   Referring now also to FIG. 2A, in a first-generation embodiment of the

9   invention, a transmitter 200 includes a modified-Chua circuit. The transmitter 200

10   generates a modulated chaotic signal responsively to an information signal 236. The

11   transmitter 200 has a primary 220 and auxiliary 237 capacitor. The auxiliary

12   capacitor 237 is selectively switched into the circuit to add selectively to the C_2

13   capacitance of the embedded Chua circuit. By switching the auxiliary capacitor 237

14   off and on, the transmitter oscillates according to a base oscillating regime and an

15   alternate oscillating regime, respectively. By controlling optoisolator 235

16   responsively to an information signal 236, an alternating pattern of chaotic

17   oscillations is generated which can be characterized as a modulation of the base

18   chaotic oscillation. This modulated chaotic pattern can be transmitted to a receiver

19   by transmitting a voltage V_1 tapped from point 242.

20   Referring now also to FIG. 2B, the modulated chaotic signal is detected by

21   a receiver 201 containing a modified-Chua circuit whose component properties are

22   chosen to insure that the receiving circuit 201 will exhibit the same oscillating

23   behavior as the base configuration (auxiliary capacitor 237 switched off) of the Chua

24   circuit of the transmitter 200. The need to match oscillating behaviors is to allow the

25   receiving circuit 201 to synchronize with the received signal 291. One way to match

26   the oscillating behaviors of the transmitting and receiving circuits 200 and 201 is to

27   match the values of the components that determine the oscillating behavior. The

28   resulting transmitted voltage V_1, output from the transmitter 200, is applied as input

29   signal 291 to the receiver 201. Note that the component values need not be matched

30   perfectly. It has been found that the receiver's α and β can differ by as much as

1 approximately 5 percent from the transmitter's without substantially affecting the

2 ability of the circuits to synchronize.

3 The use of an optoelectronic switch 235 avoids any need for output biasing.

4 Also, an optoelectronic switch 235 also adds no significant capacitance to the circuit.

5 A low output biasing and low capacitance of the switching element make it easier to

6 match the component values of the receiving and transmitting circuits 201 and 200

7 to insure synchronization. Alternatively, a reed switch or a field effect transistor

8 (FET) can be used to isolate the auxiliary capacitor 237 from the main circuit. To

9 provide a lower output biasing requirement, multiple FETs can be employed as a

10 single switch. In the embodiment of Figs. 2A and 2B, the modulated chaotic signal

11 is produced by varying the capacitance C_2 of the tank circuit 231 as described above.

12 That is, in the transmitter 200, the auxiliary capacitor 237 is isolated from the main

13 circuit by an optoelectronic switch 235, which effectively changes the capacitance

14 of capacitor 220 in FIG. 2A.

15 To modulate the transmitter 200, the capacitance C_2 of the tank circuit 231

16 is modulated by intermittently combining the capacitance of auxiliary capacitor 237

17 with that of capacitor 220. This capacitance corresponds to the capacitance of

18 capacitor 20 in the unmodified Chua circuit 101; that is, to C_2. By intermittently

19 altering this capacitance responsively to the input signal 236, the Chua circuit of the

20 transmitter 200 alternates between two different oscillating patterns.

21 The voltage signal $V_1(t)$ can be transmitted by any means desired. For

22 example, the output chaotic signal $V_1(t)$ can be used to modulate an optical carrier,

23 laser carrier, radio carrier, applied directly to a metallic (wire) interface, applied to

24 a speaker and transmitted as sound waves, or transmitted using any other mechanism.

25 The received signal can also be applied through an automatic gain control circuit (not

26 shown in this embodiment) for signal conditioning.

27 In the embodiment of FIG. 2B, the received signal can be applied through a

28 voltage follower 251, if desired for high input impedance, through a resistor 280, and

29 finally to a bridge point 281 of L-C tank circuit 261. Tank circuit 261 has an

30 inductor 248 and a capacitor 260 and can include a resistor in addition to its inherent

1   resistance. The L-C tank circuit is connected to the chaotic portion of the embedded

2   Chua oscillator of the receiver circuit 201 by a voltage follower 245. Current from

3   the L-C tank circuit 261 is applied through the voltage follower 245 and a resistor

4   265 whose resistance matches that of the transmitting circuit resistor 225. In other

5   words, the resistance R of the Chua circuits match.

6   By matching the resistance of 280 to that of resistor 225, tank circuit 261 is

7   driven or pumped by the incoming signal exactly as tank circuit 231 is pumped by

8   chaotic portion 222. When the C_2 values of the transmitter tank circuit 231 and the

9   receiving tank circuit 261 are identical (that is, when auxiliary capacitor 237 is

10   isolated from the tank circuit by opto-isolator 235), a time-varying voltage at 281

11   synchronizes and subsequently tracks that of the incoming signal 291. This

12   synchronization occurs because the incoming signal matches that at corresponding

13   point 282 of the transmitting circuit, so the environments of tank circuit 261 and tank

14   circuit 231 are the same.

15   Also, the voltage at point 281 is applied through a resistor 265 that is also

16   matched to resistor 225 so the environment of the chaotic part 262 of the receiving

17   circuit 201 is also the same as the environment of the chaotic part 222 of the

18   transmitting circuit 200. Thus, when the transmitter is oscillating about the base

19   strange attractor equilibrium points (*base* referring to the situation when the auxiliary

20   capacitor 237 is isolated from the transmitter 200 so all the circuit elements of the

21   transmitting 200 and receiving 201 circuits match), tank circuit 261 quickly goes into

22   an oscillating pattern that is in synchrony with that of the transmitter's tank circuit

23   231. When the auxiliary capacitor 237 is switched on by closing the optoisolator

24   235, the transmitter circuit 200 oscillates in a pattern that is no longer matched to that

25   of the receiver and the receiver 201 can no longer track the signal perfectly. That is,

26   the transmitting circuit 200 and the receiving circuit 201 no longer synchronize.

27   Note, the value of the combined capacitance C_2 can be varied over the range 1 μF to

28   0.015 μF, a dynamic range of 66:1.

29   As discussed above, in one embodiment, receiving circuit 201 and the base

30   configuration of the transmitting circuit 200 can be precisely matched, in terms of

1   their oscillating behavior, to insure that receiver 201 will alternately synchronize and

2   go out of synchronization responsively to the transmitter 200. That is, according to

3   this embodiment, the transmitter's and the receiver's behaviors must be substantially

4   matched for the transmitting and receiving circuits 200 and 201 to form an effective

5   communications device. Since β (See FIG. 1D) varies as the square of resistance,

6   precise resistors should be used in the transmitter 200 and the receiver 201. A

7   combination of a 1580 ohm fixed resistor 280 and a 200 ohm, 25-turn pot 265 can

8   be used for resistors 280 and 265 in receiver 201. This allows tuning of the receiver

9   resistors to obtain a precise match to those in the transmitter. Note, if component

10   values of the receiving circuit are chosen to match the values of α and β of the

11   transmitting circuit but depart more substantially from those of the receiving circuit,

12   the receiving circuit can still be driven into synchronization, but the response will not

13   be as strong.

14   Note that the receiver circuit with the synchronizing resistor added can be

15   configured with components that permit the receiver to produce a detectable pattern

16   at the detector output of a chaotic signal produced by a Chua receiver whose

17   component values do not match the transmitting circuit's precisely. A fully

18   functional communication system can be made since the receiver can produce

19   consistent output beat frequencies from the detector while the receiver circuit tries

20   to follow the input signal. The receiver therefore detects signals that are not

21   synchronized with the receiver chaotic parameters through a received process that

22   shifts the amplitude and phase of the incoming strange attractor. This mode of

23   operation produces a response over a wide range of α/β values as well as frequency

24   ranges for a single specified α/β combination. One can build a vocabulary using this

25   technique even when the chaotic signal attractor frequency is varied and α/β are

26   different. The synchronizing resistor produces this beat frequency effect at the output

27   of the detector.

28   A comparator 270 detects the alternating pattern of synchronizations and

29   desynchronizations of the receiving circuit 201. The received signal at 286 is

30   compared by comparator 270 with the voltage generated at 287. These two voltages

1   are at corresponding locations 242 and 287 in the transmitting 200 and receiving 201

2   circuits. When the transmitting 200 and receiving 201 circuits are in synchrony, that

3   is when the voltages follow the same time-dependent trajectory pattern, the

4   comparator output 290 is zero. When the transmitting 200 and receiving 201 circuits

5   are not in synchrony, that is when these voltages do not follow the same time-

6   dependent trajectory pattern, the comparator output 290 is non-zero. A waveform is

7   generated characterized by pulses representing the points where the two waveforms

8   diverge by some voltage difference determined by the sensitivity of comparator 270.

9   The sensitivity (voltage difference trigger) can be set by replacing the fixed

10   comparator 270 with a comparator having hysterisis. Of course there can be a DC

11   component, but that is ignored or filtered out of the comparator output 290.

12   The following is what occurs in the transmitter as the transmitting circuit 200

13   is modulated. When the auxiliary capacitor 237 is switched out of the circuit during

14   a first time interval, the transmitting circuit oscillates around the base (strange)

15   attractor at a particular attractor frequency. The circuit, of course, oscillates not at

16   a specific frequency, but about a set of attractors. That is, the modulated signal is

17   chaotically "smeared" over a range of frequencies around the current strange attractor

18   equilibrium point or points.

19   In the embodiment of FIGS. 2A and 2B, the equilibrium point(s) correspond

20   · to the base value of C_2, which corresponds to the capacitance of capacitor 220 alone.

21   Note that for some of the operating regimes illustrated in FIG. 1D, the equilibrium

22   point(s) can be dependent on the initial state also. When auxiliary capacitor 237 is

23   switched on by the opto-isolator 235, the transmitting circuit 200 begins oscillating

24   in a different pattern corresponding to a then-current initial state and the new value

25   of C_2 equal to the capacitance of capacitor 220 and auxiliary capacitor 237.

26   When a modulation pattern of successive actuations/deactuations is applied

27   through opto-isolator 235, the transmitting circuit 200 switches between the first

28   (base) signal and the second signal in accordance with the modulation pattern. In the

29   receiver 201, the transmitted $V_1(t)$ signal is applied at 281 and through resistor 280.

30   The receiver's L-C tank circuit 261 sees the received $V_1(t)$ applied through resistor

1   280 so that when the base-signal is transmitted, the $V_1(t)$ signal applied at 286 causes

2   the voltage at 281, $V_2'(t)$, to follow the same time-dependent trajectory as $V_2(t)$ of the

3   transmitting circuit 200. Given that $V_2'(t)$ in the receiving circuit is substantially

4   identical to $V_2(t)$, the $V_1'(t)$ in the receiving circuit follows the same time-dependent

5   trajectory as $V_1(t)$ of the transmitting circuit 200. Thus, the two circuits are

6   synchronized when the transmitting circuit is oscillating about the base strange

7   attractor(s). When $V_1(t)$ received corresponds to the base signal, $V_1(t)$ and $V_1'(t)$ are

8   substantially identical and an output 290 of comparator 270, to whose inputs $V_1(t)$

9   and $V_1'(t)$ are respectively applied, has a nominally zero amplitude. In a practical

10 system, the output of the comparator can have some hysteresis so that the difference

11 must exceed some nominal level to generate a nonzero output.

12   When optoisolator 235 switches in the auxiliary capacitor 237, transmitting

13 circuit 200 oscillates about the second strange attractor(s). When that happens, the

14 receiving circuit 201 can no longer synchronize in response to the $V_1(t)$ applied at

15 281. This is because C_2' (the capacitance of capacitor 260) no longer matches C_2 (the

16 combined capacitance of capacitor 230 and auxiliary capacitor 237). As a result, the

17 pattern of oscillation of $V_1'(t)$ is no longer synchronized with $V_1(t)$ transmitted and

18 the output 290 of comparator 270 is no longer zero. The changes in output 290

19 between the nominal zero and nominal oscillating states can be registered through

20 some convenient system, such as a power meter or frequency counter on the output

21 290, to generate a signal carrying the data in the information signal 236.

22   The receiving circuit 201 is very sensitive to the chaotic signal of $V_1(t)$. That

23 is, the incoming signal can be a very low peak power signal (power within a narrow

24 frequency band about a peak) while still cause the receiving circuit to synchronize

25 with the transmitting circuit. Thus, receiving circuit 201 is able to follow the distinct

26 chaotic time-dependent pattern of the received signal (when the base signal is being

27 transmitted) and only by virtue of its inherent tendency to follow (be driven by) it,

28 is the receiver 201 able to pick up the signal out of the noise. It is very much a

29 resonance phenomenon, even though it is not periodic in the sense of strict linear

30 oscillatory phenomena. This kind of resonance effect allows the power to be so

1   spread over a range of frequencies that the transmitted signal appears to be "buried"

2   in the noise of the channel. The signal can be detected by means of a receiving

3   circuit whose parameters are closely matched with the transmitting circuit or by

4   applying a large FFT engine (1024 point or larger) to the sampled data. The values

5   of the components of the transmitting circuit must be known in order to pick up the

6   data signal masked in the chaotic signal applied to the channel. The sensitivity of the

7   various embodiments described below is so great that an effective communications

8   signal can be characterized by a zero dB signal-to-noise ratio.

9   Note that a higher signal-to-noise ratio will allow synchronization to be

10   achieved with less precise component matching. To achieve reliable transmission

11   with the lowest signal-to-noise ratio, the precision of component-value-matching

12   should be as high as practical. To achieve precise resistance matching, as mentioned,

13   fixed resistors can be used with 25-turn fine-tuning resistors, either in parallel or

14   series for each resistor 280 and 265.

15   As discussed above, a large number of different operating regimes are

16   available by modifying C_2. However, the receiver 201 may not be capable of

17   distinguishing among these because of the inability of the receiving Chua circuit to

18   track a received signal 291 other than the one generated by the base configuration of

19   the transmitting circuit.

20   Referring now to FIGS. 3A and 3B, a transmitter 300 has a bank of auxiliary

21   capacitors 237a-237n that can be selectively switched into the tank circuit portion of

22   the transmitter 200' by respective opto-isolators 235a-235n. Opto-isolators 235a-

23   235n are controlled by a controller 201 responsively to the input signal 236. In all

24   other respects, transmitter 200' is identical to transmitter 200 of FIG. 2A.

25   To produce a modulated signal, controller 201 is programmed to receive a

26   data word at 236 and control opto-isolators 235a-235n to switch one or more of them

27   into the circuit to produce a particular chaotic pattern. For example, if opto-isolators

28   235a and 235b are actuated to switch in auxiliary capacitors 237a and 237b, the

29   capacitances of capacitors 220, 237a, and 237b are combined to produce a

30   corresponding chaotic signal. This unique pattern forms one word or symbol of a

1 vocabulary of possible ones that can be generated by actuating opto-isolators

2 according to the various possible combinations.

3 Receiver 301 includes a bank of receivers 201a-201n, each the same as shown

4 in FIG. 3B but with different component values chosen to allow each to resonate with

5 a corresponding configuration of the transmitter 200'. That is, the component values

6 for each receiver 201a-201n, are chosen such that each will synchronize with one

7 word of the vocabulary of chaotic signals generated by the transmitter 200'. The

8 outputs of each receiver 201a-201n each correspond to the output 290 of the

9 comparator of FIG. 2B. These are individually applied to a detector 271 that

10 determines which of the outputs 290a-290n is in synchrony with the transmitter 200'

11 at a given point in time. Detector 271 determines this by indicating which output

12 290a-290n is nominally equal to zero.

13 Output 290' can be a stream of digital words each corresponding to the word

14 transmitted. Thus, at any given instant, signal 290' can indicate the current, or most

15 recently received, word being applied in the received signal to input 291'. The

16 number of bits that can be transmitted with each cycle of the opto-isolators 235a-

17 235n is equal to the $\log_2$ (log base-2) of the number of different capacitor values C_2

18 forming the vocabulary. In this embodiment, the number of receivers 201a-201n

19 matches the size of the vocabulary, that is, the number of symbols or words generated

20 by the transmitter 200'.

21 Referring to FIGS. 4A and 4B, an alternative embodiment of the invention

22 also produces a vocabulary of signal-words. In this embodiment, the values of the

23 various components are chosen so that all "words" of the vocabulary are produced

24 by configurations that maintain the same α - β combinations. That is, the values of

25 C_1, C_2, R, and L of the transmitting circuit 400 are varied to produce a variety of

26 selectable chaotic oscillation patterns, each of which coincides with the same α / β

27 combinations. A bank of parallel inductors 330 can be provided that can be switched

28 in by respective opto-isolators to add to the inductance of the fixed inductor 330'. To

29 this end, a bank of parallel capacitors 320 can be provided that can be switched in by

30 respective opto-isolators to add to the capacitance of the fixed capacitor 320'. A bank

1   of parallel capacitors 315 can be provided that can be switched in by respective opto-

2   isolators to add to the capacitance of the fixed capacitor 315'. Finally, a bank of

3   parallel resistors 325 can be provided that can be switched in by respective opto-

4   isolators to change the resistance of fixed resistor 325'.

5   In the embodiment of FIGS. 4A and 4B, instead of modulating the behavior

6   of the transmitting circuit by varying only one component, the values of

7   combinations of components determining C_1, C_2, R, and L are varied in such a way

8   as to maintain $\alpha=C_2/C_1$ and $\beta=R^2C_2/L$ constant under the control of a controller 305.

9   That is, values of the auxiliary capacitors, resistors, and inductors in banks 315, 320,

10   325, and 330 are chosen so that when predefined combinations of the corresponding

11   opto-isolators are actuated, the oscillating frequency changes (See equation (1),

12   below), but α and β remain the same.

13   In a first embodiment, the inductance L and the capacitances C_1 and C_2, only,

14   are varied, maintaining R constant. That is, only the opto-isolators in banks 315,

15   320, and 330 are switched to modulate the circuit. The fundamental frequency of the

16   tank circuit, given by

17     $$F=1/[2\pi(LC_2)^{1/2}] \qquad (1)$$

18   varies even though the oscillating regime remains constant. In this embodiment,

19   there is no need to vary R to maintain a constant α and β. Also, the non-linear

20   ·resistor 310 components do not have to be varied. In one embodiment of the

21   transmitter 400, one combination of values of C_1, C_2, and L (corresponding to one

22   state of the opto-isolator switches) can be identical to the combination of values

23   corresponding to a receiving circuit 301. This combination of values causes the

24   transmitting circuit to oscillate at a base frequency corresponding to the frequency

25   of the receiver. In a first alternative embodiment, the values of C_1, C_2, and L are

26   varied, keeping R constant.

27   The receiver 401, shown in FIG. 4B, consists of a tank circuit 361 and a fast

28   Fourier transform FFT computer 379. The component values of tank circuit 361 are

29   chosen to match the α/β combinations of the transmitter 400. The value of resistor

30   380, in the first embodiment, can be chosen to match that of the fixed R value of the

1 transmitter. (As mentioned above, R is fixed in the first embodiment. The

2 embodiment depicted in FIG. 4a is capable of variable R values, but a single fixed

3 value is used in the first embodiment. If a fixed R value is to be used, the bank of

4 resistors 325 could be omitted leaving only a single resistor 325'.)

5 The output 390 of the transmitter 400 is applied (through some transmission

6 medium) to an input 391 of the receiver. The FFT block 379 determines the word

7 transmitted by the transmitter by detecting the difference between the fundamental

8 frequency, given by equation (1), of the instant configuration of the transmitter 400,

9 and the fixed configuration of the tank circuit 361. Fast Fourier transform (FFT)

10 block 379 "sees" this frequency difference because tank circuit 361 attempts to

11 follow the trajectory of the incoming signal applied through resistance 380. Since,

12 however, the incoming signal is at a different frequency, tank circuit 361 is not able

13 to maintain this trajectory and it "falls off the rails." The frequency with which the

14 tank circuit alternately follows and uncouples from the incoming signal applied at

15 391 is equal to the difference between the fundamental attractor frequency of the

16 transmitter 400 and that of the tank circuit 361. A peak at this difference frequency

17 is indicated by the output 378 of the FFT block. Thus, output 378 indicates each

18 word generated by the transmitter 400.

19 In an alternative embodiment, the value of R (the resistance determined by

20 the configuration of resistor bank 325) of the transmitter 400 is varied also. In this

21 case, the value of resistor 380 of the receiver 301 can be any of the values used for

22 R in the transmitter 400 or another value.

23 In the receiving circuit 201 of Figs. 2A and 2B, the Chua circuit oscillates in

24 synchrony with that of the transmitter 200 when the base signal is received. The

25 resulting output 290 from the comparator 270, as discussed, is a flat (nominally zero)

26 output. However, when the transmitter 200 is placed in a configuration such that it

27 a signal other than the base signal, the comparator output 290 of the receiver 201

28 becomes substantially non-zero and, also, chaotic due to the lack of synchronization.

29 This would also be the case if the transmitter 400 were connected to the receiver 201

30 of the prior embodiment. Any signal other than ones that drive the receiver into

1 synchronization will produce a non-zero chaotic signal at the output 290. In other

2 words, if this output 290 were viewed on an oscilloscope, the trace would appear

3 chaotic and it would be difficult if not impossible to tell which "word" of the

4 vocabulary of strange attractors was generating the transmitted signal. This is

5 because V_1' can no longer synchronize with the V_1 transmitted.

6   In the combination of the transmitter 400 and the receiver 301 of the FIG.

7 4A/4B embodiment, however, the transmitter drives the tank circuit 361 for part of

8 the tank circuit's fundamental cycle but subsequently skips ahead or lags behind

9 (unless, of course, the fundamental frequency of the transmitter is the same as that

10 of the receiver) the driving input voltage applied at 391. This happens at the

11 frequency difference (a beat frequency; the difference between the fundamental

12 frequency of the transmitter (e.g., (1)) and the fundamental frequency of the receiver)

13 and thus forms a detector. The reason the receiver is able to follow the transmitter

14 in this way is that the α and β of the transmitter and receiver are matched and the

15 natural frequency of the tank circuit is an integer multiple of the frequency of the

16 transmitting circuit given by equation (1).

17   In addition to using a FFT calculator, alternative ways of detecting the beat

18 frequency include an amplitude detector connected across the same terminals as the

19 FFT calculator and which outputs to a counter. The counter can count the number

20 of beats to determine the frequency of the transmitter. Various other alternatives

21 would occur to a practitioner skilled in the art of signal conditioning.

22   Referring now to Figs. 4A and 4C, an alternative receiver design employs a

23 synchronizing resistor 385 and comparator input resistors 386 and 387. In this

24 embodiment, the received signal, corresponding to a signal other than the base

25 strange attractor, is applied to the chaotic portion 362 through the synchronizing

26 resistor 385. In the embodiment of FIG. 4C, it has been found, through experiment,

27 that the synchronizing resistor 385 can be adjusted to optimize the ability of the

28 receiver circuit 402 to follow the trajectory of the transmitter signal at portions

29 thereof so that the zero levels of the output 390 are clean and clear. The application

30 of the received signal to the chaotic portion 362 of the receiver 402 through

1 synchronizing resistor 385 forces the chaotic portion into an oscillation pattern that

2 forms clear and distinct beats with the original transmitted signals that are output by

3 the comparator 370. Because the beats (amplitude differences) are distinct and clear,

4 a counting circuit 392 can be used to indicate the detected word. As in the

5 embodiment of Fig 4B, the beats indicate the frequency of the transmitting circuit

6 400 oscillations. It appears the reason the synchronizing resistor 385 allows clean

7 beats to be formed is that the receiver is forced by the application of $V_1(t)$ to the

8 chaotic portion to follow certain portions (i.e., zero crossings) of the chaotic signal

9 from the transmitter 400. During these brief intervals, the comparator output 390

10 goes to nominal zero output.

11 Counting circuit 392 can be any of various circuits for determining the

12 number of such zero-intervals per unit time. For example, the counting circuit can

13 reset a timer at the start of a nominal zero (zero within a tolerance) and output a pulse

14 to a counter if the zero is maintained for some threshold interval. The counter can

15 be reset through the transmission of a base chaotic signal for which the comparator

16 output remains zero for a second threshold interval. The counter can automatically

17 reset at predetermined intervals maintained in synchrony by the transmission of the

18 base chaotic signal.

19 Note that the use of a photo field-effect transistor (FET) optocoupler for the

20 opto-isolator in the various transmitter embodiments is recommended because of the

21 following properties. First, the example used for evaluating the circuits tested has

22 a gallium-arsenide infrared emitting diode coupled to a symmetrical bilateral silicon

23 photodetector. The detector is electrically isolated from the input signal and

24 performs as an ideal FET. Distortion free control of low level alternating current

25 (AC) and direct current (DC) signal is enabled. The primary feature that relates to

26 modulating chaos is the low output impedance of the FET when active (i.e., ≤100

27 ohms). Also, the device exhibits high impedance (≥300 Mega-ohms) when switched

28 off.

29 The output FET is controlled via a channel voltage that does not require

30 external biasing of the device so that it can operate as an ideal switch. This is

1   compared to conventional FETs that require external biasing. In addition, the device

2   is characterized by a shunt capacitance of ≤ 15 picofarads. The circuits allow

3   switching speeds of up to 2 Megahertz. For slower speeds, a solid state relay switch

4   can be used. Finally, a normal FET has some capacitance between the drain and the

5   gate. The small current differential permitted by this can prevent the circuit from

6   operating chaotically or it can make it difficult to match the chaotic pattern of the

7   base signal in the receiving and transmitting circuits.

8   The embodiment of FIG. 4D, receiver 403, has the characteristics of that of

9   FIG. 4C except the synchronizing resistor is formed of the sum of resistors 385, 387,

10   and 388. As in the embodiment of FIG. 4C, the embodiment of FIG. 4D also uses

11   a simple counter circuit for determining a beat frequency. Also, the combination of

12   resistors 385, 387, and 388 provide a synchronizing element to lock the chaotic

13   portion of the receiver to the incoming signal, as in the embodiment of FIG. 4C. In

14   this embodiment, however, a voltage follower 351 isolates the incoming

15   communication signal from the receiver-generated chaotic signal.

16   It has been determined that a resonance voltage difference is achieved when

17   the resistance of the synchronizing resistor is approximately 5 times the resistance

18   of resistor 380. This configuration maximizes the voltage difference between points

19   286 and 287 of FIG. 4D and FIG. 2B, respectively.

20   The embodiment of FIG. 4E (receiver 404) adds an emitter follower 353 to

21   isolate the oscillator portion 361 from point 287. In other words, a voltage follower

22   blocks feedback from point 287 to the oscillator portion. It provides feed-forward

23   coupling of the incoming signal 391. This causes synchronization to be forced

24   through two parallel paths. One path is through the emitter follower 351 and

25   synchronization resistors 385, 388, and 387 to the chaotic portion of the Chua circuit.

26   The other path is through the oscillator portion 361 and the emitter follower 345 and

27   resistor 365.

28   In the FIG. 4F embodiment, receiver 405 reverses the feedback path using an

29   emitter follower 363 to provide a phase lock loop type operation in which the voltage

30   V_1' is fed back into the oscillator portion of the Chua circuit to achieve

1 synchronization. In this embodiment, voltage follower 363 isolates the receiver-

2 generated signal from the incoming signal and allows the receiver generated signal

3 to feed back into the oscillator portion of the Chua circuit to cause faster

4 synchronization.

5 Referring now to FIG. 5, a general chaotic communications system in which

6 the synchronization concept identified above is applied includes a transmitter 580

7 and a receiver 590. Transmitter 580 includes a first subsystem 500 signally coupled

8 to a second subsystem 505. Subsystems 500 and 505 can include common elements,

9 but are not coextensive. First subsystem 500 applies a signal S_1 to second subsystem

10 505 and second subsystem 505 applies a signal S_2 to first subsystem 500. First

11 subsystem 500 drives second subsystem 505 with signal S_1 and second subsystem

12 505 drives first subsystem 500 with signal S2.

13 Either first subsystem 500 or second subsystem 505 can be driven by a

14 driving signal or by some current source 515 or simply by establishing an initial state

15 if both systems are lossless as in a purely numerical system generated by a computer.

16 Current source 515 could form a portion of either subsystem 500 and 505 as in a

17 Chua diode described in detail in various prior art publications, for example, Chua's

18 Circuit: A Paradigm for Chaos, Ed. Rabinder N Madan, (see pp. 13-24), World

19 Scientific Publishing, NJ, USA, 1993; the entirety of which book is incorporated

20 herein by reference. Alternatively, in a physical system such as an electrical circuit,

21 current source 515 could be derived from a driving circuit such as described in US

22 Patent No. 5,473,694 (element 115 in FIGS. 1 and 7) the entirety of which patent is

23 incorporated herein by reference. Again in a numerical model, current source 515

24 would be absent since the transmitter 580 could be considered lossless.

25 The transmitter is perturbed by some means to modulate the signal S_2 to

26 convey information. This can be done by injecting current into either signal S1 or

27 S2 or some portion of either or both of the first and second subsystems 500 and 505

28 or by modifying parameters of either or both of the first and second subsystems. This

29 information signal injection process is represented by the application of information

30 signal S_i to either of the first and second subsystems 500 and 505. Note, however,

1   that the information signal can be coextensive with a driving current source 515 such

2   as described in US Patent No. 5,473,694 and that it can be applied as a current

3   addition to either or both of signals S_1 or S_2 or by application to either of subsystems

4   500 or 505.

5   In accordance with one aspect of the invention, signal S_2 is applied to the

6   receiver 590. In the prior art arrangements, the signal S_2 would be applied only to

7   a third subsystem 530 whose configuration closely matches first subsystem 500.

8   Transmitted signal S_2 is also applied through a synchronizing filter 550 to the fourth

9   subsystem 535 whose characteristics closely match those of second subsystem 505.

10   To extract the received signal, the incoming signal S_2 is processed together with the

11   signal S_4 generated in the response, for example by subtracting them in a comparator

12   with hysteresis 540. Alternative methods of processing the received signal and the

13   signal S4 generated by the receiver can be employed, for example, the phase

14   comparison technique described in USP 5,473,694.

15   Note that in all of the above embodiments, it may be desirable periodically

16   to send a registration signal, for example, a contiguous series of base chaotic signals,

17   to insure that any clocks on the receiving and transmitting ends are aligned. Such

18   registration might be needed, for example, to demarcate the time blocks applied to

19   the FFT calculator so that the correct time series is sampled. This might not be

20   necessary, depending on the size of the FFT block relative to the symbol duration.

21   Note that the invention can be implemented using a computer rather than

22   discrete components since the Chua circuit is readily susceptible to digital

23   simulation. In this case, all the above-described processes could be implemented

24   digitally, with appropriate use of A/D and D/A conversion at either end of the

25   communications channel. The communications systems described also form the

26   basis of computer algorithms that can be implemented in a processing architecture

27   to implement a chaotic communications system. References to discrete components

28   and time-varying parameters should be interpreted to encompass their analogues in

29   the digital signal processing domain. Similarly, references herein to "circuits" should

30   be understood to include both analog and digital implementations of the circuits.

II. SECOND-GENERATION EMBODIMENTS AND TECHNIQUES

1   As described above, various first-generation systems use modulated chaotic circuits to transmit information in a low-power, high-noise environment with simple circuit elements. Modulated chaotic communication systems such as those described above provide benefits similar to spread-spectrum technology (e.g., covert communication applications and noise immunity) using a simpler, cheaper circuit design and improved signal-to-noise ratios.

The present inventors have discovered that the modulation techniques used in first-generation systems may be subject to certain bandwidth limitations. In particular, it has been determined that the modulation bandwidth using the aforementioned techniques may be generally limited to 10 to 15% of the tank circuit frequency in the transmitting circuit. This limitation is believed to be due to the fact that changing lump parameters (reactive components) in the transmitter requires a certain amount of settling time in the transmitter before the receiver can synchronize with the changed transmitter parameters, although this theory is not critical to understanding or practicing the present invention. The reactive components' impedance change is believed to cause signal spikes that ring the receiver at the discontinuity points that cause receiver noise spikes.

Modulation bandwidth is an important consideration in applying the principles of the invention commercially. In cellular telephone systems and cable television transmission systems, for example, the ability to transmit larger quantities of information over a channel of a given bandwidth translates into lower costs. Consequently, it is desirable to increase the bandwidth when modulating chaotically oscillating transmitters.

In order to overcome the aforementioned limitations, the present inventors have discovered that by modulating certain characteristics (including nonreactive components) of the transmitting circuit, the modulation bandwidth can be increased by approximately 200%. In some embodiments, this effect occurs as a result of changing one or more non-reactive resistive values in the transmitter circuit, which allows the transmitter to smoothly transition between strange attractors, which causes

1   the receiver to go into and out of synchronization almost instantly without generating

2   noise. Various transmitter circuits according to the second-generation embodiments

3   can be used with first-generation receivers described above. Other features,

4   improvements and advantages of the second-generation system will become apparent

5   through the following description and accompanying figures.

6   **A. GENERAL PRINCIPLES**

7   Turning first to FIG. 19A, a second-generation system employing various

8   inventive principles is shown. The system includes a chaotic transmitter circuit 1901

9   coupled to a communications channel 1902 and a chaotic receiver circuit 1903. Like

10 the chaotic transmitters described above with respect to the first-generation system,

11 transmitter 1901 includes a tank circuit characterized by inductor L and capacitor C2,

12 a resistor R_o, and a chaotic portion including a second capacitor C1 and a nonlinear

13 element 1904.

14 As in many of the first-generation systems, nonlinear element 1904 includes

15 a negative resistance element having a current-voltage response characteristic such

16 as that illustrated by line 1910 in FIG. 19B. In contrast to the first-generation

17 systems, however, the current-voltage characteristics of nonlinear element 1904 are

18 modulated in accordance with an information signal s(t). This modulation behavior

19 is denoted by an arrow through the nonlinear element as shown in FIG. 19A. In one

20 embodiment, nonlinear element 1904 comprises a negative resistance diode.

21 As with the first-generation systems, communications channel 1902 can

22 comprise any of various transmission media such as wire, light (including fiber

23 optic), radio frequency (all bands), or sound, for example. Chaotic receiver circuit

24 1903 can comprise any of the various receivers described above with respect to the

25 first-generation systems, or may include various enhancements described below.

26 FIG. 19B shows the current/voltage (I/V) characteristics of a nonlinear

27 element such as those used in the first-generation system. A direct current load line

28 1911 is superimposed over a piecewise linear current-voltage characteristic line 1910,

29 representing the quantity (-1/R), where R is a resistance coupling the tank circuit to

30 the chaotic portion of the circuit as shown in FIG. 19A.

1   FIG. 19C shows two single scroll attractors 1920 and 1930 orbiting around

2   equilibrium points where nonlinear element characteristic curve 1910 intersects load

3   line 1911. When operating in a single-scroll attractor mode, the current-voltage point

4   oscillates about either equilibrium ("attractor") point 1920 or 1930, depending on

5   circuit parameters. When operating in a double-scroll attractor mode, the circuit

6   transitions between upper attractor 1920 and lower attractor 1930 of the curve when

7   the strange attractor voltage touches or crosses the breakpoint voltage –Bp or +Bp

8   (i.e., a discontinuity point on the IV characteristic curve defined by the diode

9   breakdown voltage +/- Bp and any biasing voltage applied to the diodes). These

10   principles were described generally in connection with the first-generation systems

11   above.

12   According to various second-generation embodiments, the equilibrium points

13   are shifted in accordance with an information signal by changing one or more

14   resistance values of nonlinear element 1904. As shown in FIG. 19D, modulating

15   these resistance values has the effect of shifting the strange attractor points to

16   different equilibrium positions along the load line. In other words, an information

17   signal is used to shift the equilibrium points, changing the strange attractor frequency

18   and amplitude. Because this shift can be accomplished by changing a nonreactive

19   resistance value, the shift occurs quickly, instantly shifting the transmitter's operating

20   regime between different strange attractors and allowing a receiver to synchronize

21   very quickly (e.g., in less than a quarter cycle of the tank circuit's fundamental

22   frequency).

23   FIG. 19E shows the strange attractors moving with the equilibrium point,

24   wherein the impedance of the load is changing and a time varying signal is traced out

25   on the voltage across the nonlinear diode input. As shown in FIG. 19E, three or more

26   different symbols can be generated, each corresponding to a single equilibrium point

27   on the characteristic curve.

28   FIG. 19F shows one technique for changing the nonlinear diode current-

29   voltage (IV) characteristic curve using an ideal switch (SW1 and SW2) and a resistor

30   in series with the ideal switch (R5) placed in parallel with one of the nonlinear diode

1   resistors. Further details of this circuit are provided below. In general, FIG. 19F

2   shows a so-called "Caltech" diode that is modulated such that the slope Gb of the

3   nonlinear characteristic curve 1910 in FIG. 9D is changed. This causes the strange

4   attractor to change frequency and equilibrium points as shown in FIGS. 19D and

5   19E. Gb changes as follows:

6        $Gb = R_3 - R_1 / R_3 \times R_1$

7   When R5 is placed in parallel with R3 the new Gb value is:

$$Gb = \frac{\dfrac{R_3 \times R_5}{R_3 + R_5} - R_1}{\dfrac{R_3 \times R_5}{R_3 + R_5} \times R_1}$$

16   This yields a new slope for Gb and this new slope crosses load line 1911 in FIG. 19D

17   at points 2 or 3.

18   FIG. 19G shows the result of modulating the voltage across the nonlinear

19   element. Note that both the amplitude of V1 and the frequency are changed. At the

20   receiver, the frequency and/or amplitude changes can be used to demodulate the

21   signal as shown in various first-generation embodiments.

22   In general, chaotic receivers described above with reference to the first-

23   generation systems can be used to detect and recover information transmitted with

24   second-generation transmitters. As one example, the receiver shown in FIG. 4C can

25   synchronize to a transmitter and, when the transmitter's strange attractor is changed

26   in accordance with the principles described above, the receiver will fall out of

27   synchronization. These changes can be detected by a counting circuit or any of

28   various other methods described in the first-generation systems in order to recover

29   information.

30   In general, changing the current-voltage characteristics of a circuit element

31   in a chaotic transmitter by changing a nonreactive circuit value to transmit

32   information will be referred to as "chaotic trajectory shift modulation" and the

33   resulting signals will be referred to as "chaotic trajectory shifted" or "chaotic

1   trajectory modulated" signals. In the specific case where a chaotic signal is

2   transitioned between two or more discrete signal profiles by such techniques, the

3   modulation will be referred to as "chaotic trajectory shift keying modulation" and the

4   resulting signals will be referred to as "chaotic trajectory shift keyed" signals.

5   **B.**   **MODULATING A NONLINEAR ELEMENT OF THE TRANSMITTER**

6   Techniques for modulating characteristics of a nonlinear element in a chaotic

7   transmitter will now be described according to various second-generation

8   embodiments of the invention.

9   FIG. 6A shows a chaotic oscillator including a tank circuit 231, capacitor 215,

10   resistor 225, and a so-called "Kennedy" diode 600 that can be used as the nonlinear

11   element of a Chua transmitter shown in FIG. 2A. This diode, which is known in the

12   prior art (see, e.g., Chua's Circuit: A Paradigm for Chaos, Rabinder N. Madan, pp.

13   86-87 (1993)), includes op amp 602 coupled to resistors 601, 603, and 604, and op

14   amp 605 coupled to resistors 606, 607, and 608. References herein to "Kennedy

15   diode" should be understood to refer to subcircuit 600 of FIG. 6A and its many

16   obvious variations. Tank circuit 231, resistor 225, and capacitor 215 are the same as

17   or similar to those shown in the transmitter circuit of FIG. 2A.

18   In general, Kennedy diode 600 of FIG. 6A is characterized by a piecewise

19   linear current/voltage function across its terminals as illustrated in FIG. 8 by

20   segments 860a, 860b, and 860c (the entire line will be referred to as element 860).

21   The slope of segments 860a and 860c, referred to as Gb, is defined by various

22   resistive values within the nonlinear element as described below. Similarly, the slope

23   of segment 860b, referred to as Ga, is defined by resistive values within the nonlinear

24   element. The negative breakpoint –Bp and the positive breakpoint Bp, which define

25   where the slopes changes occur, are also determined by resistive values in nonlinear

26   element 600. It will be appreciated that the piecewise linear characteristics illustrated

27   in FIG. 8 represent only one of several different possible nonlinear circuits.

28   Load line 850 represents a current-voltage characteristic of a resistive element

29   coupling the oscillator to the chaotic portion of the circuit (e.g., resistor 225 in FIG.

30   2A), and is superimposed on the graph in FIG. 8 to illustrate the location of the

1 strange attractor equilibrium points. As explained previously, the intersection of load

2 line 850 with the current-voltage characteristic curve determines the location of the

3 strange attractor equilibrium points. Depending on the orientations of these curves,

4 the transmitting circuit can be made to operate in a single-scroll or double-scroll

5 attractor mode. See, e.g., FIG. 1D. Other multiple-scroll modes, such as triple-scroll

6 and the like are also possible. Single-scroll attractors orbit about the points where

7 load line 850 intersects Gb lines 860a or 860c. Double-scroll attractors transition

8 through the Ga line 860b and move to or from Gb lines 860a and 860c. A triple-scroll

9 attractor can orbit around the IV characteristic curve origin at the center of the Ga

10 860b line.

11 Various second-generation embodiments of the invention modulate a

12 nonlinear circuit element by shifting the equilibrium point of the strange attractor (for

13 single-scroll circuits) or attractors (for multi-scroll circuits). This can be done by

14 changing slopes Ga, Gb, or both, of characteristic curve 860, which can be

15 accomplished by changing various resistances within the nonlinear circuit element.

16 Consequently, one or more slopes are changed in accordance with an information

17 signal in order to modulate the position of the strange attractor (see generally FIG.

18 19D). In various first-generation embodiments, load line 850 can be modulated.

19 This load line can be used to shift the equilibrium point and change the operating

20 point as shown in FIG. 1D (element 150). Control of the load line, however, affects

21 the tank circuit reactive load, which impacts modulation performance.

22 For a single-scroll attractor, the circuit will oscillate in only one quadrant of

23 the characteristic curve (see FIG. 19C), and thus only one piece (e.g., either 860a or

24 860c, depending on which quadrant the circuit is oscillating) must be shifted to

25 accomplish this modulation. For circuits operating in a double-scroll attractor mode,

26 both pieces 860a and 860c must be simultaneously shifted to cause both attractors to

27 shift along the curve, since at any particular point in time it cannot be predicted

28 which quadrant the circuit will be operating in. In second-generation systems, one

29 can control the transition of a double-scroll or single-scroll attractor by modulating

30 either Gb 860a or Gb 860c to cause the attractor to change from the upper scroll

1 region to the lower scroll region or vice versa. In this case, only one Gb line needs

2 to be changed. This can cause a single scroll attractor to change quadrants and

3 remain in the other quadrant until its Gb line is modulated to cause a transaction.

4   FIG. 6B shows a so-called "Caltech" diode, which produces a characteristic

5 curve similar to that of the Kennedy diode. Caltech diode 650 includes two diodes

6 652 and 655 connected to a four-resistor network comprising resistors 651, 653, 654,

7 and 658, each of which is coupled to either a ground, the negative voltage supply, or

8 the positive voltage supply. Additionally, an op amp 657 and two resistors 656 and

9 659 are coupled in the circuit. As explained below with reference to FIG. 7A,

10 resistor 660, which is coupled between op amp 657 and ground, provides one

11 possible mechanism for modulation in accordance with various principles of the

12 second-generation system. References herein to "Caltech diode" should be

13 understood to refer to subcircuit 650 of FIG. 6B and its many obvious variations.

14   FIG. 6C shows yet another negative resistance diode circuit 680 that can be

15 used in a modulated transmitter according to various principles of the present

16 invention. The novel circuit shown in FIG. 6C includes four diodes 681 through 684

17 arranged in pairs with opposing polarity across the terminals of capacitor 215 through

18 resistors 685 and 686, respectively. Op amp 689 and associated resistors 688 and

19 690, along with resistor 687, perform functions similar to those of the Caltech diode

20 shown in FIG. 6B. References herein to "SAIC diode" should be understood to refer

21 to circuit 680 of FIG. 6C and its many obvious variations.

22   Turning now to FIG. 7A, one technique for modulating a nonlinear element

23 will be described in detail, using the Caltech nonlinear diode embodiment as the

24 basis for discussion. (In FIG. 7A, elements with numerals matching those in FIG.

25 6B are the same as those in FIG. 6B). As shown in FIG. 7A, Caltech diode 650 is

26 modified by adding a modulation circuit 700 including switch 735c and resistor 761

27 in parallel with resistor 660. An information signal 736 modulates switch 735c and

28 causes resistor 761 to be switched into and out of the circuit. When switch 735c is

29 open, a chaotic transmitter circuit incorporating diode 650 oscillates chaotically

30 around a single scroll or double scroll attractor. (The size of the modulating resistor

1 controls the type of scroll obtained). When switch 735c is closed, resistor 761

2 changes the resistance of the nonlinear element in the transmitter, thus changing the

3 slopes Ga and Gb (see FIG. 8) and moving the position of the strange attractor to a

4 new equilibrium position (see FIGs. 19D and 19E). This new equilibrium position

5 causes a receiver synchronized to the previous equilibrium position to fall out of

6 synchronization, thus transmitting a bit of information.

7 In accordance with various second-generation embodiments, an information

8 signal modulates a chaotic transmitter by changing a resistive part of the nonlinear

9 element in the circuit, thus causing a near instantaneous shift (less than a quarter

10 cycle of the tank circuit's fundamental period) in strange attractor characteristics in

11 the circuit. Other implementations that use transistors, field effect transistors, or

12 other active devices can effect the modulation scheme described in this paragraph

13 assuming they implement the general current-voltage characteristic curve shown in

14 FIG. 8. The nonlinear element can be modulated using different resistive

15 components as described with respect to the first-generation embodiments (e.g.,

16 optoisolator, analog switch, field effect transistor, relay, etc.) using a resistor in series

17 with an ideal switch and placing this combination in parallel with the nonlinear diode

18 resistors. Other variations of modulating the resistive value are of course possible.

19 In FIG. 6A, for example, an additional resistor can be switched in parallel

20 with resistor 608 to change the strange attractor trajectory in a Kennedy diode

21 implementation. In FIG. 6C, for example, an additional resistor can be switched in

22 parallel with resistor 687 to change the strange attractor trajectory. Other

23 combinations of resistors can also be changed in the circuits to change the trajectory

24 in any desired manner. In general, changing these resistances in the nonlinear

25 element of the chaotic transmitter circuit increases modulation bandwidth from

26 approximately 10-15% of the tank circuit fundamental frequency to 200% of the tank

27 circuit frequency when compared to modulating the reactive components.

28 Many different resistive components of the circuits described above (e.g.,

29 resistors 651, 653, 654, 658, 659, 660, etc.) can be modulated using an ideal

30 switching element to modify the trajectory in phase space of the chaotic strange

1   attractor. Good results were obtained by controlling resistor 660 of FIG. 6B to

2   simultaneously change the slope of Ga and Gb as shown in FIG. 8. As shown in

3   FIG. 7A, by switching resistor 761 into and out of the circuit, the rotation of the

4   strange attractor phase plane changes as the slope of the current-voltage characteristic

5   curve shown in FIG. 8 changes. This implementation changes the slope of both Ga

6   and Gb simultaneously and increases the modulation and demodulation rate (i.e.,

7   synchronization rate) at the various receivers described with reference to the first-

8   generation systems.

9   With reference to FIGS. 7A and 8, slope Ga is related to resistor 660 (R1) by

10   the equation:

11           $Ga = -1/R1$

12   and slope Gb is related to resistor 653 or 654 (R2) and the resistance of the switch

13   735c by the equation:

14           $Gb = -(R2-R1)/R1R2$.

15   Referring again to FIG. 8, modulating resistor R1 shifts the characteristic

16   curve of the transmitter by moving the slopes Ga and Gb. At the receiver, this

17   characteristic curve is seen as a voltage difference between the received signal and

18   the signal generated by the receiver (see, e.g., FIG. 4C). This voltage difference at

19   the receiver is independent of the type of strange attractor (i.e., single or multiple

20   scroll) and results in an improvement over transmitters that use only single-scroll

21   attractors. This modulation technique also increases the available set of strange

22   attractors, thus increasing the number of simultaneous transmitters that can operate

23   in a given environment and increasing noise immunity.

24   Returning again to FIG. 7A, frequency changes caused by modulating the

25   inductor 230, capacitors 215 and 220, or resistor 225 cause zero crossing voltages

26   (when changing from an upper to a lower scroll or vice versa) that generally cannot

27   be detected by the aforementioned receivers. This inability to detect zero crossing

28   (Ga region changes) voltage differences results in lost bits and therefore bit errors at

29   the receivers when using a double or multiple scroll attractor. Modulating the

30   nonlinear diode current-voltage characteristic curve provides an improvement in

1 detection within a Ga region that allows the use of any type of attractor as the carrier

2 for the modulating digital bit stream, rather than only a single scroll attractor. This

3 modulation technique allows modulation to be applied at two times the tank circuit

4 frequency compared to 10-15% of the tank circuit that can generally be achieved by

5 modulating the other components of the transmitter.

6 In FIG. 7B, switching element 735c in series with resistor 761 (which

7 together are in parallel across the nonlinear diode resistors) can be used to modulate

8 components 660, 653, and 654 of FIG. 7A. The resistors 653, 654, 660 can be

9 individually modulated or modulated at the same time to achieve different

10 characteristic curve shapes. Elements 653, 654, and 660 can be modulated singly to

11 change slope Gb in the upper left quadrant of FIG. 8 or Gb in the lower right

12 quadrant of FIG. 8. They can be modulated as a pair to change both Gb slopes.

13 Resistor 660 can be modulated singly to cause Gb and Ga slopes on the characteristic

14 curve to change, or it can be modulated in conjunction with resistors 653 and 654 to

15 cause larger variations in the Gb slope. The Gb slope is related to the Ga slope in the

16 Caltech diode of FIG. 7A as follows:

17 $Gb = Ga - 1/R_2$ where R_2 is element 653 or 654.

18 FIG. 7C summarizes the effect of resistor changes in nonlinear diode on slopes Gb

19 and Ga. In FIG. 7C, slope element Gb (elements 860a and 860c in FIG. 8) and Ga

20 (element 860b of FIG 8) are changed by varying the resistors for the Caltech and

21 Kennedy diodes. The Ga and Gb slopes are shown down the left hand side of the

22 page and the resistor values are shown across the top of the table. The slopes that are

23 affected by varying the resistors singly or in pairs is shown by an "x" in the box. It

24 should be noted that several resistors affect all slopes and several effect only one or

25 two slopes. Other variations beyond those specifically shown in FIG. 7C are

26 possible. Changing the values of resistors 651 and 658 (see FIG. 7A) changes the

27 characteristic curve break points (see FIG. 8) by changing the breakdown/biasing

28 points of diodes 652 and 655. The same technique can be used with any other

29 nonlinear diode characteristic curve where the components that determine the

30 nonlinear diode current-voltage characteristic curve can be modulated in a like

1   manner to produce the characteristic curve modulation.

2   The parallel resistors can be modulated over a wide range. To maintain

3   chaotic behavior, characteristic curve 860 should intersect load line 850 at three or

4   more points, as shown in FIG. 8. The load line can be right on top of Ga 860b and

5   intersect the entire Ga line as well as the two breakpoints that lie on both the Gb and

6   Ga. The closer the parallel combination of switch 735c, modulation resistor 761 and

7   original resistor 660 are to the original resistance 660 value, the less the characteristic

8   curve is changed and the more difficult it is to detect the modulation. The larger the

9   combined resistive difference between modulation resistor 761 and original resistor

10   660, the larger the voltage difference generated at the receiver (i.e., the greater the

11   characteristic curve is changed). Similar results can be obtained using the Kennedy

12   nonlinear diode of FIG. 6A by modulating the various resistors in the diode (e.g.,

13   resistors 604, 608, 607, and 601).

14   FIG. 7B shows generally how an information signal 736 can be used to make

15   and break a switch 735c (or other switch-like device) to modulate a resistance 720

16   in a nonlinear circuit element according to various second-generation embodiments.

17   The resistance modulation scheme of FIG. 7B can be applied to any type of negative

18   resistance nonlinear element, including the nonlinear diode circuits of FIG. 6A, 6B,

19   and 6C, and others. In general, a resistive value 720 in the nonlinear element is

20   modulated across its terminals 705 and 710 by applying a second resistive element

21   761 in accordance with an information signal, which may comprise for example an

22   on-off keying type signal. It will be appreciated that multiple values and switches

23   can be used to create a multi-key and multiple modulation techniques. For example,

24   multiple resistors and switches can of course be used to create a "vocabulary" of

25   resistive values, each corresponding to a different equilibrium point on the current-

26   voltage characteristic curve, with corresponding modulation and demodulation.

27   In FIG. 7B, the modulation scheme can be used to individually modulate one

28   resistor at a time or used to modulate multiple resistors simultaneously to get a

29   different strange attractor. This can be done in pairs or all resistors in FIGS. 7A, 6A,

30   6B, and 6C in any combination. This creates flexibility in strange attractor designs.

1   It is also possible to modulate resistor 720 directly using a photoresistor or FET

2   devices' internal resistance, rather than switching a resistance. Moreover, other types

3   of nonlinear circuit elements could be used in place of a diode. Examples include gas

4   breakdown tubes or operational amplifiers.

5   FIG. 9A shows various modulation limits of the nonlinear diode characteristic

6   curve used in a Chua's circuit. Starting with the load line 970 defined by -1/R, it can

7   be seen that the load line varies from -Vsat 950 to +Vsat 960. These two points are

8   the extremes of the nonlinear diode operational amplifier's voltage swing and are

9   related to the power supply voltage (Vcc) by approximately (Vcc-0.712 volts) in

10   actual practice.

11   Referring to both FIG. 9A and 9B, the nonlinear diode curve regions (901,

12   966, and 902) are also limited by this maximum voltage point. The current is limited

13   by the maximum current drive of the operational amplifier and this limit can be

14   exceeded depending on the slope of 901, 902, or 966. In addition, to operate in a

15   chaotic state, load line 970 should intersect the characteristic curve in at least three

16   points (901, 966, and 902). If these conditions are met, the transmitter will oscillate

17   chaotically because break points 971 and 940 will be crossed by the voltage in the

18   tank circuit and unpredictable behavior will take place in the form of an

19   unpredictable transition in the strange attractor orbits. As explained in more detail

20   herein, chaotic operation can also occur through the use of linear circuit elements,

21   and the invention is not limited to the use of nonlinear circuits to effect chaotic

22   operation.

23   Under the above conditions, the modulation limits for chaotic operation are

24   defined by the line segments 905, 970, and 945, and line segments 965, 966, and 945.

25   The lower curve is bounded by the -1/R load line (970) and the points where the

26   slope of the line defined the points 905 and 945 go to zero (i.e., no longer have a

27   negative slope). If the transmitter slope (905 or 945) becomes positive then a

28   receiver would have difficulty resolving where it was operating on the receivers'

29   characteristic curve since two points could have the same current (i.e., one on line

30   905 and one on line 970).

1   The second characteristic curve boundary is defined by regions 925, 965, and

2   935 and occurs where the current produced by a given slope Ga or Gb exceeds the

3   nonlinear diode's current capability, or where Gb (925) no longer touches the load

4   line curve (970), within the operational amplifiers' saturation voltage regions 950 or

5   960, or where the slope Gb (925 or 935) is zero or positive.

6   FIG. 9B shows graphically how modulation in the transmitter and receiver

7   operates. Transmitter voltage V1 (903) in the nonlinear diode produces a voltage V2

8   (904) in the tank circuit across resistor R associated with load line 935. V1 is

9   transmitted through a channel to a receiver.

10   The receiver is a decoupled version of the Chua's circuit (see, e.g., FIG. 19A).

11   The receiver's tank circuit is coupled to the channel through a resistor R that is equal

12   to R at the transmitter with a load line 931. Therefore, current flows in accordance

13   with V1 impressed on the load line 931, which generates a current I and voltage V2

14   (906). V2 (906) is coupled through a unity gain operational amplifier to the

15   nonlinear portion of the receiver, which generates a current I equal to the current

16   across the resistor R from the channel to the tank circuit. This current then produces

17   a voltage V1 (907) on the receivers' nonlinear diode characteristic curve. This

18   voltage V1 (907) can be compared with the input voltage (904), which generates a

19   voltage difference across a synchronizing resistor Rsync, (e.g., resistor 385 in FIG.

20   4C), thus generating a bit of information. When the transmitter is then switched back

21   to the fundamental characteristic curve equal to the receiver's characteristic curve,

22   the receiver and transmitter are in synchronization and a zero voltage difference

23   across Rsync 385 is produced. The incoming voltage then equals the voltage

24   generated in the receiver. The voltage difference across Rsync is therefore zero.

25   FIG. 9C provides an analysis of the positive-going breakpoint of the

26   nonlinear diode, representing the voltage points in FIG. 6B. The negative-going

27   breakpoint can be set in a similar fashion. The negative resistance 660 is generated

28   by the operational amplifier 657 in the Caltech diode shown in FIG. 6B. The voltage

29   at point 242 must overcome the breakdown voltage of diode 655 and the voltage bias

30   set up by voltage V+ set by the voltage division network made up of resistors 658

1   and 654. This ratio is defined as (V+)*{[654]/([658] + [654])}. The sum of the

2   voltage of resistor 655 and (V+)*{[654]/([658] + [654])} sets the breakpoint for the

3   nonlinear diode.

4   There is another way to shape the characteristic curve and set the chaotic

5   operation. These same resistors 658, 654, and 660 can also set the breakpoint. The

6   diode voltage breakpoint 655 can also be raised by adding diodes in series as shown

7   in FIG. 6C. Changing 660, 653, and/or 654 change the slope of the nonlinear load

8   line. This same technique can be applied with any nonlinear diode. These diodes

9   will have a different set of resistor values that have to be manipulated in a like

10   fashion to the Kennedy diode of FIG. 6A. Other types of circuit elements could be

11   used to modulate the nonlinear elements, such as photoresistors, FETs or other types

12   of transistors could be substituted for the resistors to vary the resistance of these

13   elements.

14   The zero crossings of the transmitted and received signal are maintained

15   irrespective of the curve being used at the transmitter. However, there is a difference

16   voltage at all other points on the receiver curve when the transmitter is using the

17   modulating curve.

18   The equivalent resistance of the modulating resistor and the resistor being

19   modulated is given as follows:

20   .   Parallel resistance equivalent = {(resistance of switch [735c] +

21   [761])*([660])}/ (resistance of switch [735c] + [660] + [761]) (Equation 2).

22   This equation can be used to set any desired modulation resistor size to

23   achieve the voltage difference to meet a particular communications systems noise

24   performance requirements.

25   FIG. 10 shows a field effect transistor 1001 placed across diodes 652 and 655

26   to implement on-off keying according to a second-generation embodiment. In this

27   embodiment, FET 1001 implements on-off keying, which allows rapid switching but

28   results in a detectable modulation frequency in the voltage output of the Chua's

29   circuit across C1 (255) or C2 (260) in FIG. 2B. However, it is an improvement over

30   current practice, which results in a better detection at the receiver output across a

1 comparator detector.

2 Reference will now be made to FIGS. 20A through 23I, which show various

3 data plots and displays depicting voltages, currents, and frequencies in systems

4 employing the principles of the invention.

5 Double-Scroll Attractors

6 FIG. 20A shows a voltage (V1) versus voltage (V2) versus time (T) plot of

7 a chaotic signal (double scroll strange attractor) without modulation. FIG. 20B

8 shows the plot of FIG. 20A when modulated with an information signal according

9 to various second-generation embodiments. Voltage V1 is taken at point 242 in a

10 chaotic transmitter (voltage across capacitor 215, see FIG. 7A), while voltage V2 is

11 taken at point 282 in the same transmitter (voltage across capacitor 220). FIG. 20C

12 shows a plot of voltage (V2) versus current (I3) in inductor 230 as a function of time

13 without modulation, while FIG. 20D shows the same quantities when modulated with

14 an information signal. FIG. 20E shows a plot of voltage (V1) versus current (I3)

15 without modulation, and FIG. 20F shows the same quantities when modulated.

16 In FIGS. 20A, 20C, and 20E, a 5.6KHz strange attractor without modulation

17 oscillates about two equilibrium points. FIGS. 20B, 20D, and 20F show modulation

18 at 1000 bps applied to the transmitter nonlinear diode slopes Ga and Gb. This

19 illustrates how the strange attractor parameters change with time and how modulating

20 the non-linear diode affects the behavior of the system as the Gb line crosses the load

21 line at various points caused by the Gb slope changing.

22 The modulated signals were modulated at 1000 bps with a bit stream of

23 "1010". The modulated signal transmits the same strange attractor for a binary "1"

24 and different strange attractor for a binary "0". The transmission of a binary "0"

25 shows a different time varying signal when compared with the modulated strange

26 attractor signal. Notice the un-modulated strange attractor signal and the modulated

27 strange attractor signal in FIGS. 20A and 20B are the same from time t = 0 to 1

28 millisecond due to transmission of a binary "1", (i.e. synchronized case) while from

29 time t = 2 to 3 milliseconds in FIGS. 20A and 20B the strange attractors differ due

30 to the transmission of binary "0", (i.e. non-synchronized case). (The same holds true

1   for FIGS. 20C and 20D, and for FIGS. 20E and 20F.)

2   The strange attractor signals are correlated whenever a binary "1" is

3   transmitted and uncorralated whenever a binary "0" is transmitted given the receiver

4   is tuned to receive a binary "1". At the receiver, the synchronized strange attractor

5   produces a voltage across the synchronizing resistor equal to zero. This establishes

6   the synchronized state and produces a series of pulses corresponding to the out-of-

7   synchronization state that corresponds to the spirals of the strange attractor projected

8   onto the V1 axis.

9   FIGS. 21A through 21C show modulation of the characteristic current-voltage

10   curve for a nonlinear diode, and FIGS. 21D through 21F show corresponding

11   voltage-current plots for the modulated transmitter corresponding to FIGS. 21A

12   through 21C.

13   In FIG. 21A, resistor R1 has a value of 1200 ohms, resulting in intersections

14   between the load line and diode curve at approximately –5 and +5 volts. In FIG.

15   21B, resistor R1 has a value of 1210 ohms, resulting in a slightly smaller voltage

16   point for the intersections (i.e., about –4.75 and +4.75 volts). In FIG. 21C, resistor

17   R1 has a value of 1220 ohms, resulting in yet a smaller voltage point for the

18   intersections (i.e., about –4.5 and +4.5 volts, respectively). Resistor R1 affects the

19   nonlinear diode slopes Ga and Gb as shown in these figures. As R1 increases, the

20   slope of Ga and Gb decreases. As Ga and Gb change, the break point moves, causing

21   the voltage swing to increase or decrease.

22   Voltage-current plots corresponding to the above modulations are depicted

23   in FIGS. 21D through 21F. FIG. 21D shows voltage V2 versus current in the

24   inductor I3 plotted against voltage V1 for the case where R1=1200 ohms. FIG. 21E

25   shows the same quantities for the case where R1=1210 ohms. FIG. 21F shows the

26   same quantities for the case where R1=1220 ohms. As can be seen in FIGS. 21D

27   through 21F, as the resistance is increased, the strange attractor trajectories become

28   "squashed." As R1 increases, the amplitude of the attractor decreases in V1, V2, and

29   I3 due to a voltage decrease. This is due to the shift in the equilibrium points around

30   the load line defined by the resistor R between the tank circuit and the nonlinear

1   diode circuit. As R1 increases, the plots rotate in a clockwise direction.

2   FIGS. 21G through 21I show frequency versus amplitude plots corresponding

3   to the modulated parameters of FIGS. 21A through 21C. In other words, FIG. 21G

4   shows the spectrum when R1=1200 ohms; FIG. 21H shows the spectrum when

5   R1=1210 ohms; and FIG. 21I shows the spectrum when R1=1220 ohms. As can be

6   seen, the frequency response shifts with an increase in R1. This frequency shift can

7   be used to determine which value of R1 is switched in the circuit. In addition, the

8   magnitude of the right most frequency spike changes with a change in R1, and this

9   amplitude can be detected using a filter and frequency detector. When R1 is set to

10   1200 ohms, 1210 ohms, and 1220 ohms, respectively, the FFT spike is at 6.6kHz,

11   6.85 kHz, and 6.95 kHz as shown in these figures.

12   <u>Single-Scroll Attractors</u>

13   FIG. 22A shows a voltage (V1) versus voltage (V2) versus time (T) plot of

14   a chaotic signal (single scroll strange attractor) without modulation. FIG. 22B shows

15   the plot of FIG. 22A when modulated with an information signal according to

16   various second-generation embodiments. These voltages are taken at the same points

17   as in FIGS. 21A through 21C.

18   FIG. 22C shows a plot of voltage (V2) versus current (I3) in inductor 230

19   (FIG. 7A) as a function of time without modulation, while FIG. 22D shows the same

20   quantities when modulated with an information signal. FIG. 22E shows a plot of

21   voltage (V1) versus current (I3) without modulation, and FIG. 22F shows the same

22   quantities when modulated.

23   These figures also show the effect of starting the system with a set of initial

24   conditions that are out of phase with the strange attractor. The system spirals into the

25   single scroll strange attractor and then builds to its chaotic state conditions. As in

26   FIGS. 20B, 20D and 20F, the signals in FIGS. 21B, 21D and 21Fwere modulated at

27   1000 bps with a bit stream of "1010". As with the double scroll strange attractor, the

28   strange attractor signals are correlated whenever a binary "1" is transmitted and

29   uncorrelated whenever a binary "0" is transmitted given the receiver is tuned to

30   receive a binary "1".

1   At the receiver, the synchronized strange attractor produces a voltage across

2   the synchronizing resistor equal to zero. This establishes the synchronized state and

3   produces a series of pulses equating to the out-of-synchronization state that

4   correspond to the spirals of the strange attractor projected onto the V1 axis. The

5   single scroll attractor detection is simplified since there is no zero crossing that can

6   cause a detection error as the modulating frequency is increased.

7   FIGS. 23A through 23C show nonlinear diode plots similar to those in FIGS.

8   21A through 21C, but for a single-scroll attractor, wherein resistor R1 is 1930 ohms,

9   1940 ohms, and 1950 ohms, respectively. Resistor R1 affects the nonlinear diode

10 slopes Ga and Gb as shown. As R1 increases, the slope of Ga and Gb decreases. As

11 Ga and Gb change, the break point moves, causing the voltage swing to increase or

12 decrease.

13   FIGS. 23D through 23F show corresponding phase plots for the different

14 resistive values in FIGS. 23A through 23C. These phase plots show that as R1

15 increases, the amplitude of the attractor decreases in V1, V2, and I3 due to a voltage

16 decrease. This is due to the shift in the equilibrium points around the load line

17 defined by the resistor R between the tank circuit and the nonlinear diode circuit.

18 The phase plots also show that as R1 increases, the plots rotate in a clockwise

19 direction.

20   FIGS. 23G through 23I show frequency versus amplitude plots corresponding

21 to the different resistive values in FIGS. 23A through 23C. The plots show how the

22 frequency response increases with an increase in R1. One can use this shift in

23 frequency to determine which value of R1 is switched into the nonlinear diode

24 circuit. In addition, the magnitude of the right most frequency spike changes with

25 a change in R1 in the nonlinear diode circuit, and this amplitude change can be

26 detected using, for example, an amplitude detection circuit. At R1 values of 1930,

27 1940, and 1950 ohms respectively, the FFT spike is at 6.425 kHz, 6.375 kHz, and

28 6.350kHz, respectively as shown.

29   To summarize, a receiver that receives a signal generated in accordance with

30 the above modulation techniques would see a variation of the current-voltage

1   characteristic curve of the incoming signal and therefore a difference between the

2   incoming signal and the signal generated by a matched chaotic circuit at the receiver.

3   Second-generation modulation techniques change the trajectory of a component that

4   varies the trajectory about the strange attractor. This shows up in the signal as an

5   equivalent (or apparent) amplitude variation defined by the voltage across component

6   215, 220 and the current through the inductor 230 in FIG. 2A. This is different from

7   conventional phase shift keying where the signal phase is shifted. In this case, the

8   trajectory of the strange attractor determines a current through inductor 230, and the

9   two voltages across capacitors 220 and 215 are changed within phase space when

10   modulation is applied to one of the resistors in the diodes shown in FIGS. 6A or 6B

11   using an ideal switch (analog, FET, or relay) to switch a resistor in parallel with any

12   nonlinear diode resistor components (e.g., any of resistors 601, 603, 604, 606, 607,

13   608 in FIG. 6A, or any of resistors 651, 653, 654, 658, 659, or 660 in FIG. 6B), or

14   any combinations thereof.

15   Shifting the chaotic oscillator strange attractor trajectory by changing the

16   current-voltage characteristic of the transmitter's nonlinear diode results in an ability

17   to modulate the strange attractor trajectory at modulation rates as high as

18   approximately two times the tank circuit fundamental frequency (f_{LC}) (the Nyquist

19   rate $1/(2f)$ given by the tank circuit components) and still recover the signal using any

20   of the receivers described in the first-generation embodiments. This is an increase

21   of the modulating data rate by approximately 200% over using first-generation

22   transmitter modulation techniques that change a frequency by modulating other

23   reactive circuit components (e.g., elements 230, 215, or 220 in FIG. 2A) or resistor

24   225.

25   Modulating Ga and Gb causes the receiver to synchronize at the Nyquist rate

26   of $1/(2f)$. The output pulses exhibits this rise time at the receive output of a

27   comparator or subtractor circuit. This technique can also be applied to a tunnel diode

28   nonlinear current voltage characteristic curve for radio frequency operation or to a

29   laser amplifier configured as a nonlinear diode.

30   The Fast Fourier Transform (FFT) of the voltage across C1 (element 215) and

1   C2 (element 220) showed a 30 dB higher attenuation of the modulation frequency

2   component which was lost in the chaotic signal noise floor when modulating the

3   nonlinear diode using R1 (element 660 in FIG. 7A) by less than one ohm. The

4   modulating frequency was therefore not detectable as a distinct frequency component

5   since it was below the chaotic frequency spectrum noise floor.

6   Modulating the current-voltage characteristic curve results in a random

7   polarity reversal of the attractors, and at the receiver demodulated output where the

8   detected data is recovered. This random reversal adds a significant improvement in

9   the detectors' ability to recover an information signal in secure communication

10   environments since it helps mask the information signal in what appears to be

11   additional noise on the channel.

12   A similar effect can be achieved by modulation components 653 and 654

13   since only Gb is modified. It will be appreciated that modulation can be effected by

14   changing the breakpoints in order to shift equilibrium positions on the characteristic

15   curve.

16   **C. <u>RECEIVER SYNCHRONIZATION</u>**

17   The following description explains how the chaotic receiver synchronizes

18   with signals generated using the techniques and systems described above, and how

19   the value of the synchronizing resistor Rsync can be selected for optimal

20   performance in a chaotic transmitter implementation.

21   Referring back to the receiver embodiment of FIG. 4C, an analysis of

22   synchronizing resistor Rsync (element 385) when the nonlinear diode is modulated

23   determines the maximum voltage difference with and without modulation applied

24   across that resistor. As Rsync increases in value, the voltage difference across it

25   increases with modulation as τ (Tau) = 1/(resistor 385 x capacitor 355) = 1/(Rsync

26   x C_1). This is an exponential time decaying voltage function. It has been determined

27   that the decay rate of the time function τ optimizes the synchronizing signal

28   difference as Rsync approaches zero ohms but the voltage difference across resistor

29   385 approaches zero. The upper synchronization voltage difference between a

30   synchronizing signal and a modulating signal occurs at Rsync = ∞ (infinity).

1   In practice, Rsync can be optimized. It was determined empirically that the

2   maximum frequency that could be demodulated in a receiver was two times the tank

3   circuit (element 361) resonant frequency f_{LC}, since a modulation rate greater than 2

4   f_{LC} at capacitor 360 in the tank circuit charges the capacitor and does not allow

5   sufficient time for the capacitor to discharge. Capacitors 360 and 355 then charge

6   to one of the equilibrium points of the strange attractor. On the characteristic curve,

7   voltage V1 goes to a value equal to the crossing of the direct current load line (1/R)

8   where it crosses Gb in the upper or lower quadrant as shown in FIG. 9B (line 904).

9   Since this is a limiting modulation rate, then this point can be used to set the time

10   constant for the $1/(Rsync \times C_1)$ time constant. This permits a modulation rate of $2f_{LC}$

11   and allows the largest voltage difference across the Rsync resistor while still

12   obtaining a synchronization time that supports optimum detection of the amplitude

13   and frequency difference of the incoming signal (i.e., optimizes the voltage

14   difference across Rsync versus synchronization time). One can reduce Rsync even

15   further but the voltage difference is then exponentially decreased across Rsync for

16   both the synchronized and modulated signal. Optimum synchronization (in a

17   noiseless channel) for circuit design can therefore be obtained when Rsync is selected

18   as follows:

19   Rsync $\leq (1/(2f_{LC} \times C_1))$ where C_1 is capacitor 355.

20   This is an optimum design point for Rsync in a chaotic receiver. The

21   frequency f_{LC} can be determined by taking the fast Fourier transform of the

22   transmitter or by applying the following formula for an RLC tank circuit:

23   Frequency $= k(1/(3.41(L \times C_2)^{.5})) = k(1/(3.41(L \times [\text{element } 360])^{0.5}))$

24   where k is a constant ranging from 1 to approximately 2.5 depending on the internal

25   resistance of the inductor (element 348) and the resistor R (element 380).

26   This establishes the upper limit for Rsync. It has been determined

27   experimentally that in the presence of noise, Rsync in the range of less than about 1

28   ohm to about 1000 ohms improves the signal-to-noise ratio of the receiver circuits

29   in the first and second generation receivers by several (e.g. two or more) orders of

30   magnitude. This Rsync coupling resistor value forces the nonlinear diode portion of

1 the circuit to synchronize faster. The synchronization time in a noisy channel

2 dominates the system's noisy performance. Thus, Rsync can be adjusted to optimize

3 the energy per bit to the noise per Hertz of bandwidth. This can be achieved at

4 resistance values for Rsync from about 1 ohm or less to Rsync = $(1/(2 \, f_{LC} x \, C_1))$, see

5 FIG. 4C, depending on the noise characteristics of the channel.

D. GAIN CONTROL AMPLIFIER

7 The chaotic receivers exemplified in FIGS. 4C, 4D, 4E, and 4F can be

8 optimized by adding an amplifier in the circuit. FIG. 11 shows a gain control

9 amplifier 1146 inserted into a receiving circuit that reduces or eliminates the need for

10 an automatic gain control (AGC) amplifier on the input at point 1191. As shown,

11 gain control amplifier 1146 is inserted into the circuit between elements 1145 and

12 1165. Resistors 1188 and 1189 set the amplification. In one embodiment, the

13 optimum gain is 2.4 dB to 3 dB.

14 The amplifier will also operate with gains above 3 dB. However, the size of

15 the gain should be adjusted to ensure that the signal is not clipped. This amplifier

16 reduces the need for an automatic gain control (AGC) amplifier on the input of the

17 chaotic receivers at point 1191. With amplifier 1146 in the circuit, the incoming

18 signal can be attenuated by more than 10 dB and chaotic synchronization will still

19 occur in subsystem 1126 to generate voltage V1 across capacitor 1155. This

20 amplifier also seems to provide some noise immunity because of its band pass filter

21 characteristics and its ability to move the signal away from the breakpoint of the

22 nonlinear diode element 1126.

23 The most general case for this amplifier is shown in FIG. 12, which

24 represents an improvement over the system shown in FIG. 5. In FIG. 12, amplifier

25 1201 is inserted between subsystems 1230 and 1235 to overcome signal attenuation

26 in the incoming channel. An extension of this is to add an amplifier between every

27 subsystem that performs the function of a receiver element, such as subsystem 1230

28 and subsystem 1235. The elements in FIG. 12 can be roughly mapped to previously

29 described embodiments as follows: first subsystem 1200 corresponds to the tank

30 circuit in the transmitter; second subsystem 1205 corresponds to the nonlinear

1 element in the transmitter; third subsystem 1230 corresponds to the tank circuit in the

2 receiver; fourth subsystem 1235 corresponds to the nonlinear element in the receiver;

3 and synchronizing subsystem 1250 corresponds to the synchronizing resistor and

4 associated components. The voltage across tank circuit capacitor 1160 in the

5 receiver's tank circuit is amplified prior to being applied to resistor 1165.

6 The IV characteristic curve in FIG. 9B illustrates how this amplifier works.

7 The received signal is compared to the signal produced by the receiver. If the

8 received signal was produced on Gb slope 901 or 902 in the transmitter and it is

9 attenuated below the breakpoint of the receiver 971, then it is being reproduced based

10 on the slope of Ga 966. The voltage in the tank circuit must be amplified enough to

11 raise it above breakpoint 971. This is based on slope Gb and load line 972. This

12 then causes a voltage on the nonlinear diode Gb slope 925, or 931 instead of Ga 966.

13 If the new voltage is compared to the incoming signal, there will be a constant

14 difference between the two signals. This then brings the systems into

15 synchronization, which can be detected as a one or zero voltage by various threshold

16 detection circuits discussed herein.

17 In summary, the amplifier reduces the need for a precise automatic gain

18 control in the receiver system of a chaotic communications system designed from

19 Chua's circuit or any other nonlinear set of decoupled dynamical nonlinear equations

20 used as a receiver. The amplifier allows the input channel signal to be attenuated up

21 to 10 dB or more. This means that the automatic gain control does not have to

22 precisely match the incoming signal. Moreover, the amplifier also acts as a noise

23 canceling circuit based on its bandwidth filtering characteristics.

24 E. CHAOTIC SIGNAL FILTERING TRANSMITTER

25 Certain second-generation embodiments of the present invention also include

26 filtering in both the transmitter and receiver circuits. If a signal is produced by the

27 transmitter and filtered, then the receiver can still follow this signal. At the receiver,

28 the receiver follows the filtered signal and oscillates in such a manner that it actually

29 adds the missing frequency components back into the receiver generated signal. The

30 signal generated in the receiver must then be filtered using a like filter to match the

1   received signal. If the signal is not filtered at the transmitter, then it can be filtered

2   at the receiver to reduce noise effects. This can be done at the front end of the

3   receiver or after the receiver has reconstructed the signal. The following paragraphs

4   describe this process.

5   Turning to FIG. 13, transmitting circuit 1300 is similar to the transmitter of

6   FIG. 2A, but includes an amplifier 1308 that isolates filter 1309 from the chaotic

7   subsystem 222. This prevents the filter's input impedance from interfering with the

8   chaotic operation of the chaotic circuit's nonlinear resistor 284 or capacitor 215 (see

9   FIG. 2A). Filter 1309 can be a low pass filter to eliminate the direct current

10   component, or a bandpass filter to allow only the higher order chaotic signal

11   components to pass over a real channel such as a radio baseband where direct current

12   components are blocked. In practice, the three receiver designs described above

13   with respect to the first-generation systems can be synchronized with output signal

14   1343 from filter 1309. This is an improvement because it can be used to match a

15   chaotic transmitter circuit to the bandwidth of an audio, video, radio frequency, or

16   light (e.g., laser) channel, where frequencies below some cutoff frequency cannot be

17   passed in a real system.

18   Considering a bandpass filter, filter 1309 bandwidth in practice can be

19   centered around the tank circuit frequency (e.g., $\pm$ 20% of the tank circuit frequency)

20   or centered around the information rate and still synchronize at the receiver. All the

21   modulation techniques of the first-generation and second-generation embodiments

22   can be applied with the transmitter filtering principles of FIG. 13. Another way to

23   judge the bandpass requirement is to look at the bandpass of the tank circuit and

24   design a bandpass filter that is slightly wider than it by approximately 5%. Optimum

25   selection should, of course, be made on an application-dependent design basis.

26   FIGS. 14A and 14B show low pass and bandpass filter characteristics of one

27   embodiment. Points 1400, 1420, 1440 are the 3 dB points of the filters. In actual

28   practice, the bandwidth can be twice this width since modulation rates of two times

29   the tank circuit frequency are possible. Point 1430 is the center frequency of the

30   bandpass filter and corresponds to the fundamental frequency of the tank circuit 231

1   of FIG. 13.

2   In general, placing a bandpass or lowpass filter at the output of the chaotic

3   transmitter matches the transmitter to audio, video, light, or radio frequency channel

4   modulators/upconverters.   It also reduces noise components outside these

5   bandwidths.

6   **F. <u>RECEIVER FILTERING</u>**

7   As shown in FIGS. 15 through 17, chaotic receivers can also benefit from

8   noise filters. FIG. 15 shows a receiver including noise filters 1593, 1594 and 1597

9   to filter out noise components introduced by the transmission channel according to

10   a second-generation embodiment of the invention.   FIG. 16 shows a receiver

11   including noise filters 1666, 1694 and 1697 to filter out noise components introduced

12   by the transmission channel according to a second-generation embodiment of the

13   invention. FIG. 17 shows a receiver including filters 1793, 1766, 1794, 1797 and

14   1721, wherein filter 1721 works with automatic gain control amplifier 1746 to further

15   reduce noise generated by subsystem 1726. The noise filters can be lowpass filters

16   or bandpass filters that match the transmitter filter, or a combination of low pass and

17 · bandpass filters.

18   Referring first to FIG. 15, two bandpass filters 1593 and 1594 filter out noise

19   components introduced by the channel. The detector is then placed across the

20   buffered input signal after filter 1594 and receiver reconstructed signal after filter

21   1597.   This reduces the noise components and improves detection in a noisy

22   environment. The bandwidth of these filters can be set just like the description of

23   FIG. 13 and FIGS. 14A and 14B.

24   FIG. 16 provides another implementation of noise filters. In this case, the

25   incoming signal 1691 is applied to subsystem 1661 without filtering. The filtering

26   is done in the synchronizing resistor chain made up of components 1666, 1653, 1687

27   and 1694. The noise is filtered from the input and also from the output through 1696

28   and 1697.

29   The foregoing techniques can be used in conjunction with the modulation

30   scheme of FIGS. 7A and 7B to optimize the detection of the chaotic strange attractor

1 trajectory shift keying to improve a communications receiver detection performance.

2 FIG. 17 shows a receiver including filters 1793, 1766, 1794, 1797, and 1721.

3 Filter 1721 can be designed to work with the automatic gain control amplifier 1746

4 to further reduce noise generated in subsystem 1761. This receiver filtering

5 technique will work with the receiver designs described above with reference to the

6 first-generation systems. Unity gain operational amplifier 1722 acts as an impedance

7 buffer.

G. APPLICATION TO CELLULAR PHONE SYSTEMS

9 The principles of both the first-generation and second-generation systems can

10 be applied to cellular and non-cellular telephony in several different ways. First, they

11 can be used in a baseband modem. Second, they can be applied directly to an

12 intermediate frequency level modem. Third, they can be used at the radio frequency

13 level. In a cellular telephone application, for example, a large number of matched

14 strange attractor pairs could be created to transmit information, much like code

15 division multiplex systems operate today. Thus, for example, each of a plurality of

16 modulators (or transmitters) in a base station would have strange attractor parameters

17 matched to a corresponding one of a plurality of portable telephone demodulators (or

18 receivers); each corresponding demodulator/receiver in the base station would also

19 be matched to a corresponding transmitter/modulator in one of the plurality of

20 portable telephones. If implemented using digital signal processing techniques, the

21 matching can be implemented with a software change rather than requiring specially

22 matched hardware pairs.

23 In a baseband modem implementation, a modem takes advantage of the large

24 embedded base of cellular telephony equipment and infrastructure. In this

25 implementation, the cellular phone technology acts as the carrier system. The

26 advantage of this method is that chaotic synchronization techniques could be used

27 to produce digital modems for use in the current generation of cellular telephones.

28 Cellular phones would take advantage of the chaotic synchronization to synchronize

29 in low signal to noise ratio environments. This would allow digital data to be passed

30 at two to three times the current data rate of current cellular systems.

1 Turning first to FIG. 18A, various first and second-generation embodiments
2 discussed in this patent application will be applied as described. A chaotic baseband
3 modem 1801 includes a chaotic transmitter 1801A and a chaotic receiver 1801B.
4 These circuits are coupled to an interface circuit 1803, which in turn is coupled to a
5 cellular telephone 1801 (which may comprise a conventional cellular telephone of
6 various types commercially used as of the filing date of this application). In this
7 embodiment, the output of the interface circuit 1803 would be in the audio frequency
8 range; the chaotic modem would essentially be interfaced as though the cellular
9 phone was a radio system that accepted an audio input. Cellular telephone 1802
10 communicates with a conventional telephone system through one or more radio
11 frequency communication nodes or base stations in a conventional manner.

12 The chaotic transmitter modem frequencies would fall into the audio
13 baseband of a cellular phone system (e.g., 300-3000 Hz). An improvement in data
14 rate would result from applying the M-ary signaling techniques discussed in the M-
15 ary section below. The signal-to-noise ratio improvement over current state-of-the-
16 art modems (frequency shift keying) will allow cellular phones to operate over a
17 longer range than current systems. In other respects, the current generation cellular
18 phones would remain the same.

19 Turning to FIG. 18B, in an intermediate frequency implementation, a chaotic
20 modem could have its frequency adjusted to operate, for example, on a 1 MHz to 10
21 MHz intermediate frequency (IF) band, but the invention is not limited to this
22 exemplary IF range. In this family of embodiments, a chaotic intermediate frequency
23 modem 1807 comprises a chaotic transmitter 1807A and a chaotic receiver 1807B,
24 wherein the bandwidth of the intermediate frequency could be set to 1.2MHz to 2
25 MHz to match the code division multiple access signals used in current technology.
26 The radio frequency operation of the cellular telephone 1808 would remain the same
27 as today's state-of-the-art systems. This embodiment would require some changes
28 to the cellular telephone from its existing commercial implementation.

29 FIG. 18C shows a chaotic modem that operates at radio frequencies. The
30 system includes a chaotic modem 1813 including a chaotic RF transmitter 1813A and

1　a chaotic RF receiver 1813B, which is coupled to an RF interface circuit 1814 and

2　antenna 1815. In this system, the frequency of the chaotic system is increased to the

3　cellular telephone frequency band and the chaotic transmitter and receivers discussed

4　herein are operated at radio frequencies. The components may change to radio

5　frequency components such as tunnel diodes for the non-linear diode implementation

6　and radio frequency amplifiers. In other respects, the chaotic transmitting and

7　receiving circuits remain the same as in the embodiments described herein. The

8　specification for frequency use can be set to the code division multiple access

9　standard of 1.2-2 MHz. This means that the tank circuit bandpass filter would be set

10　to this 3dB bandpass and the radio frequency transmission will be filtered to reduce

11　frequency components outside of the passband.

12　　　The embodiments described above can take advantage of the characteristics

13　of chaotic synchronization signals to synchronize in low signal-to-noise ratio

14　environments. Various tradeoffs trades include (a) reducing the transmitter while

15　keeping the data rate constant; (b) reducing the number of cells for a given area; (c)

16　increasing the number of users in a given cell size; and (d) providing new digital

17　services in the same bandwidth as current cellular phone systems. The interface

18　circuits can be constructed as needed to interface existing or future telephone systems

19　using well-known techniques.

20　　　In the embodiments exemplified by FIG. 18C, the new chaotic radio

21　frequency modulation scheme could take advantage of chaotic synchronization

22　spread spectrum capability where many more chaotic synchronization signals could

23　be mixed over the radio frequency spectrum and decoded much like code division

24　multiple access is applied today. The number of users would be determined by the

25　number of orthogonal spreading codes (chaotic synchronization signals) possible and

26　the frequency division spacing characteristics of chaotic signals as discussed with

27　respect to the first-generation embodiments. Chaotic synchronization offers the

28　advantage of simpler transmitters and receivers as well as noise immunity

29　unachievable in conventional state of the art systems.

H. DUAL TRANSMITTER/RECEIVER COMBINATIONS

1

FIG. 24 shows a dual-transmitter configuration (1200, 1205) according to a
second-generation embodiment of the invention. In FIG. 24, two systems 1200 and
1205 are used to generate two strange attractor pairs (e.g., two single or two double
scroll pairs). The strange attractors can be generated using any of the techniques
described above with respect to the first-generation or second-generation systems.
These strange attractors are buffered using two unity gain operational amplifier
circuits 1201 and 1202 to prevent the switches 1220 and 1210 from loading the
chaotic circuit systems 1200 and 1205. For the purposes of explaining the operation,
the nonlinear diode subsections 251 and 1251 will form the basis for tuning the two
chaotic systems. Each transmitter 1200 and 1205 can be implementing using an
embodiment of FIG. 6A, FIG. 6B, FIG. 6C, or other embodiments described herein.

The slopes Ga and Gb of the nonlinear diodes can be set to two different
values. Chaotic circuit 1205 can include components set to different resistance
values than equivalent components of the nonlinear diodes in chaotic system 1200.
This leaves the tank circuit subsystem elements 200 and capacitor 215 in both Chua's
circuit systems with equal elements. In the non linear diodes the operational
amplifier 223 and its biasing resistors 202 and 203 are also equal. The only
difference between the two systems 1200 and 1205 is the slopes of the nonlinear
diodes 251 and 1251.

An ideal switching system 1265 including single pole double throw switches
(or similar devices)1210, 1220, inverter 1206, and information input source 1206
provide a means to switch the transmitter between the two strange attractors. A
summing circuit 1295 sums the switched output at point 1208 to provide a
transmitted signal based on the ratio of resistors 1270 and 1280 to resistor 1290 as
in any summing circuit implementation. In this case, a 1:1 ratio is used. This signal
has high frequency components due to the switching action of 1265.

To remove these unwanted switching pulses, low pass filter 1310 is used. Its
lowpass cutoff frequency is set to the highest frequency of the channel roll off. An
attenuator circuit 1360 is then added to adjust the output signal to match the signal

1   to the channel. A system such as that shown in FIG. 24 (i.e., using two switched

2   transmitters) provides more precise control of the strange attractors when compared

3   to switching a single transmitter's nonlinear diode or other elements values. Using

4   two or more switched transmitters allows one to choose the position of the strange

5   attractors in phase space (double scroll, single scroll in upper quadrant or lower

6   quadrant for digital signals). Using two or more switched transmitters provides faster

7   switching time since the switching time is no longer dependent on the reactive

8   components of the transmitter to stabilize. The switched transmitter in combination

9   with a receiver pair allows the strange attractors to be optimized for the best detection

10 results at the receiver.

11 **I. DUAL RECEIVER DESIGN USING COMMON TANK CIRCUIT**

12   FIG. 25 shows an embodiment including a dual receiver (1370 and 601) using

13 a common tank circuit 361. This design is dependent on using a dual transmitter, or

14 a single transmitter where the nonlinear diode and/or the capacitor 215 is varied and

15 the tank circuit 200 remains the same. Lowpass filter 393 reduces the noise

16 component and is set to the channel bandwidth. The unity gain operational amplifier

17 351 buffers the line and drives the multiple circuit elements. The common tank

18 circuit element consists of elements 380, 348, and 360. The dual receiver elements

19 consist of subsystem 601 and 1370. Receiver subsystems 601 and 1370 are tuned

20 to the transmitters' two strange attractor nonlinear diodes 251 and 1251 (FIG. 24)

21 plus the capacitors 215 as shown in FIG. 24. In FIG. 25, synchronizing resistors

22 1430 and 1450 provide a means to detect the difference between the incoming

23 waveform and the strange attractor produced by the dual receiver elements. Since

24 the strange attractors are different than those of the other of the subsystems 601 and

25 1370, the difference across the two resistors is either zero when matched or some

26 voltage difference when they are not matched. Unity gain operational amplifiers

27 1400, 1410, 1420, and 345 isolate the synchronizing resistors and the non linear

28 diode elements from the tank circuit and input signal 286 to prevent feedback from

29 causing the system to detune. This dual receiver provides a precise method to detect

30 two or more strange attractors. The nonlinear portions 1370 and 601 can be

1   duplicated to detect additional chaotic strange attractors.

2   A dual receiver design such as the one depicted in FIG. 25 reduces the

3 number of parts required in a receiver when capacitor 215 or the nonlinear diode are

4 modulated or when one switches between two or more transmitter circuits with a

5 common tank circuit in FIG. 24. The voltage difference across the synchronizing

6 resistors 1450 and 1430 can be maximized for detection by amplitude detection

7 circuits as discussed herein. This also provides a convenient way to detect frequency

8 variations between the two circuits at points 1470 and 287. The voltage difference

9 between points 1450 and 1430 can be maximized for amplitude modulated signal

10 detection by maximizing the Ga and Gb differences. The dual receiver circuit design

11 can be extended to N receivers where elements 345, 365, 355, and 350 are

12 reproduced with the values of the transmitter systems.

13 **J. SUBTRACTOR CIRCUIT**

14   FIG. 26 shows a subtraction circuit block for detecting the voltage difference

15 across the synchronization resistors 1400 and 1450 in FIG. 25 and to detect the

16 voltage difference between points 1470 and 287 of FIG. 25. It includes buffer unity

17 gain operational amplifiers 1710 and 1720 in subsystem 1930 which isolates the

18 subtractor circuit from the receiver circuits. This is followed by an attenuator circuit

19 subsystem 1940 consisting of elements 1730, 1750, and 1760 as well as elements

20 1740, 1745, and 1770.

21   The attenuation circuit reduces the voltage across the synchronization

22 resistors to a level the subtraction subsystems 1950 can process without clipping the

23 input signals prior to the subtraction. The subtraction circuit makes mathematical

24 detection possible in an analog system.

25 **K. ABSOLUTE VALUE CIRCUIT**

26   FIG. 27 shows an absolute value circuit made from an operational amplifier

27 circuit. The absolute value circuit can be used to take the absolute value of the input

28 signal. This circuit can be used for a detector. This circuit executes the following

29 equation:

30   Vout = abs(Vin)

1   Subsystem 1965 is an inverting operational amplifier that inverts the input

2   1880 and provides it as an output 1920 to one side of the full wave rectifier

3   subsystem 1965. The input 1880 to the inverter subsystem 1960 is input 1885 of the

4   full wave rectifier. The final output 2045 of the circuit is then further processed to

5   recover the amplitude modulated information signal shown in FIG. 24 input data

6   1206 to complete the transmitter receiver pair. This absolute value circuit allows

7   amplitude modulation on a chaotic signal to be processed into the information-

8   bearing signal of FIG. 24 input point 1206. This circuit can be replaced by a

9   squaring circuit that performs the absolute value function.

10   **L. DUAL RECEIVER SYNCHRONIZATION DETECTOR CIRCUIT**

11   FIG. 28 shows a dual receiver synchronization detector circuit. In this circuit

12   the inputs to the circuit 2000, 2010 and 2005, 2015 are placed across the

13   synchronization resistors 1450 and 1430 respectively in FIG. 25. Input 2000 is tied

14   to point 1440 and 2010 is tied to point 1470. The detector input 2005 is tied to point

15   1415. The detector input 2015 is tied to point 287. The output of each subtraction

16   circuit is then subtracted as follows: 2020 − 2030 = 2035. Then 2035 is passed

17   through an absolute value subsystem 2040 (FIG. 27). The output 2050 is then

18   processed through a moving average detector and the original modulating signal at

19   point 1206 in FIG. 24 is recovered after squaring the signal.

20   This detector provides a significant improvement in detector output over a

21   simple comparator. The detector subtracts noise off the input signal since the noise

22   across the two synchronizing resistors of FIG. 25 are subtracted off.

23   **M. DUAL RECEIVER SYNC DETECTOR USING ABSOLUTE VALUES**

24   FIG. 29 works like FIG. 28 except the output of the initial subtractions 2020

25   and 2030 are processed by an absolute value circuit to provide the absolute value at

26   2060 and 2070. These two numbers are then subtracted. The difference should have

27   the maximum difference since the dual receiver circuit in FIG. 25 alternates between

28   the two strange attractors and produce minimum and maximum values in step with

29   the modulating signals. The output of the circuit 2080 is then processed using a

30   moving average and square law detector and the original modulating signal 1208 in

1 FIG. 24 is recovered by the detector. This detector provides a significant

2 improvement in detector output over a simple comparator. The detector subtracts

3 noise off the input signal since the noise across the two synchronizing resistors of

4 FIG. 25 are subtracted off.

5 **N. SINGLE DETECTOR CIRCUIT**

6 FIG. 30 shows a single circuit detection circuit element. This element can be

7 used to process the voltage at points 287 and 1470 of FIG. 25. Good results were

8 obtained when the strange attractors generated by the dual transmitter in FIG. 23

9 were single scroll attractors with one attractor produced on the upper quadrant Gb

10 and the other single scroll attractor produced on the Gb in the negative quadrant.

11 This maximizes the difference between the equilibrium points. The circuit is

12 believed to work best when the transmitter is switched between two single scroll

13 attractors in two different quadrants of the IV characteristic curve in FIG. 9A,

14 although double scroll and single scrolls in the same quadrant would work, one

15 produced on the +Gb slope and the other on the –Gb slope.

16 **O. INTERFACTING TO COMMUNICATIONS SYSTEMS**

17 FIG. 31 shows a chaotic system interface in a chaotic circuit that allows one

18 to tap off the chaotic signal without interfering with the chaotic circuit's operation.

19 Unity gain operational amplifier 2300 provides the isolation. Since the chaotic signal

20 voltage and direct current offset at the input 2200 are not at the level a

21 communication system expects to see at the baseband input 2350 (e.g., microphone

22 input, data port input, etc.) a direct current offset circuit made up of a direct current

23 power supply 2230 and an attenuator subsection 2480 made up of one or more

24 attenuator circuits are required.

25 The use of more than one attenuator is usually required since the input to

26 most communication systems is in the range of hundreds of millivolts and the output

27 of a chaotic circuit is in the range of volts. The dynamic range of one attenuator is

28 usually not adequate and therefore several are required to match the input voltage

29 level 2200 the required output voltage level 2350. In FIG. 31 there are three

30 attenuators. The first attenuator consists of elements 2210, 2240, and 2700. The

1   second attenuator consists of elements 2290, 2300, and 2310. The third attenuator

2   consists of elements 2320, 2330, and 2340. These three attenuators make up the

3   attenuator subsection 2480. In addition to attenuators this subsection can also act as

4   an amplifier section by changing the ratio of the resistor elements 2210, and 2240 or

5   2290 and 2300 or 2320 and 2330. This makes subsection 2480 a universal

6   amplifier/attenuation element. The level shifter subsection 2470 consists of a power

7   supply 2230 and an interface resistor 2250. This allows one to add or subtract a

8   direct current level from the input chaotic signal 2200 to match a single ended

9   baseband communications system at the output 2350.

10   This may be important in modem applications since many existing radios and

11   cellular phones have single ended baseband input/output requirements. This is an

12   improvement over current practice since it eliminates the need for a direct current

13   filter and preserves the direct current characteristics at the receiver since a known

14   level can then be added or subtracted from the incoming signal. This technique can

15   be applied to an amplitude modulated communication system (e.g., audio, radio

16   frequency, or laser) and to a frequency modulated radio system.

17   The direct current offset subsystem 2470 is an improvement over current

18   practice since it eliminates the need for a direct current high pass filter and preserves

19   the direct current characteristics of the signal at the receiver where this known direct

20   .current voltage level can then be added or subtracted from the incoming signal. The

21   attenuator/amplifier subsection 2480 is an improvement in practice in interfacing a

22   chaotic signal voltage to the baseband of an existing communication system. The

23   transmitter baseband interface circuit can be used in both radio and cable systems.

24   FIG. 32 shows a chaotic receiver baseband interface circuit. The output of

25   a communications receiver system baseband can be applied to the input of the

26   interface circuit at point 2650. Normally this is a single ended output and the signal

27   must be amplified from millivolt levels to the voltage levels expected by the chaotic

28   receiver circuit at input point 2830. Following a unity gain operational amplifier

29   2660 the signal is amplified by a low noise high gain amplifier subsection 2840

30   which is followed by a unity gain operational amplifier 2770 for isolation from the

1 final gain stage consisting of elements 2780, 2820 and 2870. Inserted into the final

2 amplifier section is the direct current voltage offset subsection 2850. The output of

3 this circuit 2830 is the recovered chaotic voltage with the direct current voltage offset

4 reinserted. The overall system element is designated subsystem 2860 and becomes

5 an interface block for any chaotic receiver block.

6 The direct current offset subsystem 2850 is an improvement over current

7 practice since it adds the direct current characteristics of the signal back into the

8 receiver signal where this known direct current voltage level was subtracted at the

9 transmitter and can be reinjected into the signal for processing by the chaotic

10 receiver. The attenuator/amplifier subsection 2840 is an improvement in practice in

11 interfacing a chaotic signal voltage to the baseband of an existing communications

12 system. The receiver baseband interface circuit can be used in both radio and cable

13 systems.

14 FIG. 33 shows a chaotic interface system 2490 and 2860 used to interface a

15 chaotic transmitter and receiver to an infrared amplitude modulated subsystem 3090.

16 The infrared transmitter consists of elements 3000, 3100, 3020, 3040, infrared diode

17 3050, transistor 3060, and 3070. The subsystem 2490 shifts the direct current

18 voltage level of the input signal and attenuates it from volts of signal 2200 to

19 millivolts 2350 of signal (FIG. 31). The infrared receiver is a self biased

20 phototransistor pair 3080. The output 2650 is then applied to the chaotic interface

21 circuit 2860 at interface point 2650. The infrared transmitter and receiver pair is a

22 bandlimited system. This embodiment shows that a chaotic signal can be sent over

23 a band limited amplitude modulated system and the original signal recovered

24 adequately in the chaotic receiver to pass information bits at the channel baseband

25 bandwidth. The chaotic interface circuit subsystem 2860 (FIG. 32) then changes the

26 chaotic signal at interface point 2650 from millivolts to volts at interface point 2930

27 with the direct current voltage offset added back into the system. One infrared

28 system embodiment had a bandwidth of 30 kilohertz and the chaotic signal was a 5.6

29 kilohertz strange attractor. Applications of this system include remote control units

30 as well as communications systems. In addition to infrared systems, the inventive

1 principles can be applied to other amplitude modulated systems including audio,

2 radio frequency, and light (e.g., light emitting diodes and laser systems).

3 Band limited communications systems such as an infrared communications

4 system can support a chaotic signal modem provided the interface circuits 2490 and

5 2860 are used to buffer the chaotic signal. The interface blocks provide a

6 communications system capable of transmitting and receiving a band limited chaotic

7 signal.

8 FIG. 34 is an embodiment of the chaotic signal interface circuit to a radio

9 baseband system 3120. The resistor element 3100 provides an impedance match to

10 the input of the transmitter baseband 2350. The radio frequency transmitter 3100 can

11 be any radio transmitter at any frequency. The radio receiver 3110 is matched to the

12 communications transmitter. The output of the receiver 3110 is applied across a

13 resistor 3120. This produces a voltage at interface point 2650. The chaotic receiver

14 interface circuit 2860 then matches the receiver signal to the chaotic receiver circuit.

15 Chaotic signals can be passed over band limited frequency modulated systems using

16 the interface subsections 2490 and 2860.

17 FIG. 35 shows a cable driver system to pass a chaotic signal down a twisted

18 pair or coaxial cable system. The chaotic interface circuit 2490 buffers the signal by

19 providing attenuation and direct current voltage offset. The signal is then sent to a

20 balanced line driver subsystem 2455. The balanced line driver prepares the chaotic

21 signal to go over a twisted pair cable and matches the impedance of the cable to the

22 driver circuit. One possible cable driver is derived from an operational amplifier

23 book (SAMS "IC Op-Amp Cookbook", Third Edition, Walter G. Jung, 86-60253

24 (1997), page 387 titled differential line driver). This circuit allows chaotic signals

25 to be sent over cable systems.

26 A chaotic interface subsystem 2490 provides an advancement in practice over

27 current interface circuits and allows existing differential line drivers to be used for

28 chaotic systems.

29 FIG. 36 shows one embodiment of a cable line receiver circuit derived from

30 the SAMS Operational Amplifier book (page 346) with a chaotic receiver interface

1   circuit 2860. The subsystem 2880 buffers the signal received from the differential

2   input amplifier with high common mode rejection 2880. The circuit is a combination

3   inverting attenuator and scaling adder which rejects common-mode input

4   components while amplifying differential ones. This system allows chaotic signals

5   to be matched to a transmission line such as standard twisted pair. The   chaotic

6   interface circuit allows chaotic signals to be passed over a transmission line.

7   FIG. 37A shows an embodiment of modulating the transmitter using the

8   resistor 660 from FIG. 6B and measuring the voltage difference across element 1185

9   of FIG. 11 at the chaotic receiver. This embodiment illustrates that the transmitter

10   and receiver pair can act as a linear analog system in regions 2100. This means that

11   amplitude modulation of the nonlinear diode is possible and the modulating signal

12   can be recovered by examining the voltage difference across 1185 of FIG. 11. In

13   addition, this figure shows that there are a number of values for element 660 in FIG.

14   6B 2110, 2120, 2130, 2140, and 2150 that are usable as signal vectors or code words

15   in digital systems. Each of these resistor values generates a unique difference at the

16   receiver when the receiver is set to a fixed resistor value R1. The signal vectors can

17   be selected such that a maximum ratio is obtained. This curve can be drawn for

18   different transmitter receiver pairs for different 660 values and different parameters

19   of the Chua's circuit R, L, C1, and C2. The output of each of the receivers can be

20   correlated to a specific code word. FIG. 37A can also be generated for resistor 653

21   and 654 in parallel with 660.

22   A separate set of curves was also implemented for the Kennedy diode as

23   shown in FIG. 6A when the transmitter element 608 was modulated and the receiver

24   signal difference 3215 across elements 1450 and 1430 (i.e., Rsync) of FIG. 25 was

25   measured and plotted as FIG. 37B. The curves 3220, 3230, 3240, and 3250 are

26   examples of the voltage difference across Rsync 1450 and 1430 as resistor 228 of the

27   transmitter is varied. The linear regions of the curves (e.g., 3255) can act as an

28   analog signal demodulation curve by varying element 228 in the transmitter and

29   obtaining a linear signal output at the receiver. Note the chaotic region goes from

30   approximately point 3310 to 3320. Within this band of resistance, there are N

1   possible signal vectors where N is determined by the minimum detectable voltage

2   difference at the receiver for various resistor values 228 at the transmitter.

3   In FIG. 37B, the minimum voltage difference is where the transmitter and

4   receiver have the same nonlinear diode characteristic curves 3270, 3280, 3290, and

5   3300 (i.e., element 608 in the transmit and receive nonlinear diodes are the same).

6   Depending on the noise in the channel the number of usable vectors is reduced. For

7   example, if point 3260 is the noise floor then the vectors would have to be adjusted

8   to insure a detectable voltage difference based on the modulated value of resistor

9   608. One could also look at a fixed resistor value such as 75 ohms point 3330 at the

10   receiver and look for the voltage difference at the receiver 3315 as resistor 608 in the

11   transmitter changes between the vectors to determine a detection threshold. For

12   digital signals, this means that one can space the vectors at one ohm increments and

13   detect the voltage difference to all other signal vectors. In the presence of noise one

14   may only be able to get some portion of this number based on the voltage difference

15   trigger setting 3260 or 3315. This curve can be plotted for other circuit elements and

16   expand the number of signal vectors.

17   FIG. 37C shows what happens when the transmitter capacitor 215 is varied

18   in FIG. 6B and receiver capacitors 355 and 1490 in FIG. 25 are set to fixed values

19   with the nonlinear diode characteristic curve set at a fixed value. The voltage

20   difference across Rsync 1430 and 1450 of FIG. 25 is plotted against the transmitter

21   capacitor 215 of FIG. 6B. This same curve can be plotted for every resistor value in

22   FIG. 37B and a surface constructed showing where the optimum capacitor and

23   resistor values are for signal vectors in a digital communications system. This

24   increases the number of signal vectors possible with a given tank circuit and

25   capacitor 215. In FIG. 37C, the curve generated by varying capacitor C1 is plotted

26   as 3400. There are two linear regions of the curve defined as 3410 and 3420. These

27   regions can be used as analog modulation regions in an analog communications

28   system. The voltage differences in conjunction with FIG. 37B can be used to form

29   sets of curves that can be used to build code words when each value of R1 (608) is

30   used to generate a new set of C1 (215) curves.

1   In a transmitter receiver pair configured as a communications system, analog

2   communication is possible by modulating R1 (608) or C1 (215) in the nonlinear

3   diode of the chaotic transmitter circuit and observing the voltage difference across

4   Rsync in the receiver circuit using the linear regions of the detection curves in FIG.

5   37A, 37B, and 37C. In the chaotic transmitter and receiver pair a large number of

6   signal vectors can be constructed using the curves in FIGS. 37 A, 37B, and 37C.

7   **P. USING MULTIPLE TRANSMITTERS AND RECEIVERS**

8   Various systems and techniques described above use a technique of

9   modulating current-voltage characteristics of a circuit element to achieve enhanced

10   data rates in a basic digital communications channel. This technique can approach

11   the Nyquist rate in its signal modulation. This is a fundamental improvement over

12   the 15-20% performance of previous chaotic communications systems and is the

13   basis for further improvement to increase the information-signaling rate well beyond

14   the Nyquist rate. This is based on two fundamental factors that have been described

15   herein. First, the chaotic trajectory phase shift keying technique has been shown to

16   provide a robust method of signaling near the Nyquist rate using modulated chaotic

17   attractors, which can be detected by matched receivers using a robust process based

18   on the receiver improvements that are part of the technique.

19   Second, the transmitter and receiver circuits have been shown to be capable

20   of generating a very large diversity of chaotic attractor modulated waveforms that

21   can be discriminated from each other using the improved receiver techniques. This

22   enables a coding scheme in which each transmitted bit (or Tbit) is coded to a digital

23   word (a sequence of information bits – or Ibits) that is uniquely associated with one

24   of several transmitter/receiver/attractor combinations implemented in a multiple

25   transmitter/receiver system. Thus Tbits are transmitted as described above at a rate

26   near the Nyquist rate, thus staying within the limits imposed by basic physics and

27   information theory, while Ibits are transmitted at a higher rate, which is a multiple

28   of the Tbit transmission rate and can apparently exceed the fundamental limits of

29   information theory.

1   This technique is illustrated in FIG. 38 for the specific case in which two Ibits

2   4000 are coded into each Tbit 4065. This case requires 4 transmitters 4020, 4030,

3   4040, and 4050 and receivers 4090, 4100, 4110, and 4120 in matched combinations.

4   Each receiver is capable of uniquely synchronizing with the waveform of its matched

5   transmitter, and will not synchronize with the waveforms of the other transmitters.

6   The transmitters and receivers could be designed with identical tank circuits and if

7   values of the Ga and Gb parameters are varied as described in the chaotic trajectory

8   phase shift keying technique (see FIGS. 24 and 25). Alternately, entirely different

9   attractor types could be used for each transmitter/receiver pair so long as the

10   synchronization advantages are retained. Unlike chaotic trajectory phase shift keying

11   as described elsewhere herein, this technique is implemented using unmodulated

12   chaotic circuits. Individual Tbit waveforms are formed by switching between the

13   unmodulated outputs of the transmitters at the Tbit signaling rate. This has the same

14   effect and the same characteristics as the previous modulation technique but is more

15   efficient when several or a large number of transmitter/receiver pairs are to be

16   implemented.

17   The received signal consists of a series of Tbits, each with the characteristic

18   waveform transmitted by the transmitter that corresponds to its coded information

19   content. The train of transmitted Tbits consists of a patched sequence of different

20   waveforms, and is to be analyzed by the receiver circuit to recover the original

21   sequence of Tbits in which the information signal is encoded. The signal is fed

22   synchronously to all the receivers, which implement a demodulation and detection

23   process similar to that described herein. The receiver that matches the transmitter of

24   a given Tbit will synchronize using the rapid synchronization methods described

25   herein, while the other receivers will not. In each Tbit period the matched receiver

26   is identified and the coded word is extracted to recover the original information

27   signal, or the Ibits in the coded word.

28   FIG. 38 starts with element 4000 for example, the four signal coding vectors.

29   The input information sequence 4005 controls the order in which the coded 2 bit

30   sequence vectors are transmitted via a set switch 4010. This set switch then controls

1   the order in which the transmitters are passed through the channel via switch 4060.

2   The Tbit sequence 4065 is then injected into the RF transmitter 4070 using a chaotic

3   interface circuit such as that of FIG. 31 (element 2490). The RF receiver 4080 then

4   receives the signal and processes it to the baseband 4105 where it is applied to a

5   chaotic receiver interface circuit such as that shown in FIG. 38 and processed to

6   match the receiver input impedance the chaotic receivers 4090, 4100, 4110, and

7   4120. Each of the transmitters can be a chaotic circuit (such as a Chua's circuit)

8   tuned to a specific strange attractor. After the chaotic receivers are detector circuits

9   4130, 4140, 4150, and 4160. Each of the receivers can be a matched Chua's circuit

10 receiver as discussed herein. The detector outputs are sent to selector switch logic

11 4160 that declares the received vector 4170. The code is then recorded in a record

12 code buffer and recovered in a 2-bit sequence buffer to be played out in a serial

13 fashion as the recovered bit sequence 4200.

14   The example in FIG. 38 uses a coded word consisting of 2 Ibits. The 4

15 combinations of Ibit values result in a total of 4 digital words. With 4

16 transmitter/receiver pairs each word is assigned uniquely to one pair so that a

17 complete coding (transform) of Ibits to Tbits is accomplished. Assuming that the

18 basic communication process for Tbits (which follows the processes described

19 previously) and that the process achieves a Tbit signaling rate of S bits/second with

20 a bit error rate of BER at Tbit energy E_b/N_0, the word-coded process would achieve

21 twice the true information transmission rate at half the equivalent (per Ibit) E_b/N_0 and

22 the same BER. Thus the extension of the basic chaotic trajectory phase shift keying

23 technique doubles the information rate without fundamentally changing the

24 communications process.

25   This enhancement process is generalized by observing that 2^N

26 transmitter/receiver pairs could be used to provide a complete code for digital words

27 of N Ibits each. Thus a 2^N dimension system has the capacity to increase information

28 rate by a factor of N without fundamentally changing the structure and physical

29 communications parameters of the chaotic trajectory phase shift keying system

30 described elsewhere herein. If N is large then information data rates (Ibit signaling

1 rates) can be achieved far beyond Shannon's limit for communications systems.

2 Although there are other methods that exceed Shannon's limit in transmission rate,

3 this technique has the fundamental advantage that it does not require the deep

4 transmitted signal modulation that can limit the compatibility of other techniques

5 with various types of communications media with their requirements for sharing

6 communications spectra.

7 This technique may require the use of error correcting codes. Although

8 increasing information rate by a factor of N does not increase the basic bit error rate,

9 each Tbit error results in N correlated Ibit errors, analogous to a frame error in other

10 digital communications systems. This is a disadvantage that must be addressed by

11 error coding. However, the technique has the desirable property that the error

12 correction codes will operate on Tbits, at their lower transmission rate, in a

13 conventional way. No new error correction techniques will be required to achieve

14 acceptable performance with this technique despite the very high equivalent Ibit

15 transmission rate that can be achieved. Various techniques for interfacing the

16 trajectory shift keying signals to real communications systems at audio, radio

17 frequency and laser light frequencies are discussed herein. These techniques translate

18 to baseband modems that can be used in cable transmission, radio and laser systems.

19 This technique can be applied to a wide variety of systems using analog, hybrid

20 digital-analog or pure digital components for implementing the chaotic circuits. The

21 structure of such a system is shown in FIG. 39. Thus a system that uses a smaller

22 number of transmitter/receiver pairs (up to approximately $2^8 = 256$ pairs) could use

23 low cost ASIC chips with analog or analog and digital components for both

24 transmitter and receiver systems. Much higher data rate systems (exceeding $2^{12} =$

25 4096 pairs) could be implemented in custom CMOS or other integrated circuit chips

26 to provide a system with very high performance.

27 FIG. 39 is the general case where a codebook is made up of "N-bit" code

28 blocks 5000 with 2^N Transmitters 5030 and receivers 5090. The transmitters are

29 multiplexed by element 5040 and switched in based on the input bit sequence 5105

30 by 5020 and 5060 giving "Tbits" 5065. The signal is then interfaced to the RF

1 transmitter 5070. The RF receiver 5080 is interfaced to the 2^N chaotic receivers

2 5090. The output of the chaotic receivers are then multiplexed 5100 and decoded

3 using 5110. The 2^N vectors are then recorded 5120 and recovered 5130. Finally the

4 data is serialized into the recovered data stream 5140 to become the recovered bit

5 stream 5140.

6   For purposes of reference, the following is an estimate of the enhanced

7 performance that can be achieved with this technique. Shannon's limit implies a

8 limit of 2 bits per second (BPS) per Hertz (Hz) of channel width. The basic chaotic

9 trajectory phase shift keying technique could operate at a Tbit signaling efficiency

10 of about 80% with error correction coding reducing Tbit efficiency by about 60% to

11 a net efficiency of 32%. This would yield a Tbit signaling rate of .64 BPS/Hz of

12 channel width. Without the enhancement this system could transmit 9600 BPS

13 through a typical 15,000 Hz channel. A 2^8 dimension system would increase this to

14 76,800 BPS through the channel, which is far beyond the capabilities of other

15 existing systems. A 2^{12} dimension system could increase this to 115,200 BPS.

16 Performance at these levels would allow a more redundant level of error coding

17 (multiple techniques or coding plus redundant transmissions) while maintaining high

18 performance. Using this technique information data rate is limited only by the

19 physical ability to fabricate components with a high packing density of transmitter

20 and receiver circuits, and by availability and design of switching components to

21 implement the process at the chip level.

22   Combined with the basic techniques for modulating and demodulating chaotic

23 signals described herein, the multiple transmitter/receiver enhancement applies to all

24 communications applications envisioned for this technology. The technique

25 comprises a breakthrough for applications that demand high information (Ibit) data

26 rates but have fundamental limitations on physical or Tbit transmission rates. These

27 include restrictions on use of the communications spectrum and limitations due to

28 transmission effects in the communications medium itself. Thus very high speed

29 modems will be possible using this technique, so long as the basic communication

1   process supports the use of basic chaotic trajectory phase shift keying at conventional

2   Tbit signaling rates as described elsewhere herein.

3   Although there are other methods that exceed Shannon's limit in transmission

4   rate, this technique has the fundamental advantage that it does not require the deep

5   transmitted signal modulation that can limit the compatibility of other techniques

6   with various types of communications media with their requirements for sharing

7   communications spectra. The technique comprises a breakthrough for applications

8   that demand high information (Ibit) data rates but have fundamental limitations on

9   physical or Tbit transmission rates. These include restrictions on use of the

10   communications spectrum and limitations due to transmission effects in the

11   communications medium itself.

12   Very high speed modems are possible using this technique, so long as the

13   basic communication process supports the use of basic chaotic trajectory phase shift

14   keying at conventional Tbit signaling rates as described elsewhere herein. The same

15   technique can be applied to the Gb-only transmitter and receiver pair discussed below

16   by modulating the equilibrium points (e.g., by shifting to M-ary equilibrium points).

17   **Q. DIGITAL VERSION WITH CUBE-LAW NONLINEAR COMPONENT**

18   Various chaotic trajectory phase shift keying techniques described herein use

19   a chaotic oscillator circuit that has been modified in several respects to improve

20   performance, but which retains the nonlinear diode component (for example, element

21   650 in FIG. 6B) in the nonlinear diode circuitry. When an all-digital implementation

22   of the chaotic trajectory phase shift keying techniques is desired, however this

23   component can be implemented more efficiently by other means. One digital

24   implementation of chaotic trajectory phase shift keying can use a block diagram

25   component using the following equation to model the current vs. voltage

26   characteristic of the nonlinear diode:

27 $$I = -aV - bV^3$$

28   where I is the current through the nonlinear diode and V is the voltage across the

29   diode. The constants a and b are calculated from the G_a, G_b, and B_p characteristics

30   of the nonlinear diode circuit. The constants are constrained by two conditions

1 necessary for the component to produce the chaotic modulations necessary to

2 implement the communications technique. These conditions are as follows:

3 Condition 1: the slope of the equation at I=0, V=0 must be less than –Ga so that the

4 curve passes through the load line and will produce chaotic modulations, producing

5 the constraint that a > G_a.

6 Condition 2: the maximum usable value of V is defined by the point at which the

7 slope of the I vs. V curve reverses sign and becomes unusable, producing the

8 constraint that the maximum desired voltage V_{max} is as follows:

9 $$|V_{max}| \leq (-a/3b)^{1/2}.$$

10 The third requirement that the curve passes through the origin at V=0 and I=0 is

11 automatically satisfied by this form of the equation.

12 The constants a and b can be calculated by imposing further constraints on

13 the shape of the curve. Good performance can be obtained by using the following

14 procedure: (1) set V_{max} at the actual maximum voltage consistent with the circuitry

15 or other components used to fabricate the chaos producing circuit, and (2) constrain

16 the curve to pass through a particular point (V*,I*). These produce enough

17 information to solve for the constants. As an example, if the curve is constrained to

18 pass through the breakpoint in the nonlinear diode (B_p,-G_aB_p) then the coefficients

19 can be solved as follows:

20 Let: $r = (B_p/V_{max})$ and $r \leq 1$;

21 Then: $a = 3G_a / (3 - r^2)$;

22 $b = - (G_a/V_{max}^2) / (3 – r^2)$.

23 Other solutions can be derived by changing the particular point (V*,I*) through

24 which the curve is constrained to pass.

25 In a digital implementation of the equation it could be calculated as I = -

26 V*(a+b*V*V), requiring the floating-point calculation of three multiplications, one

27 addition, and one sign change.

28 This nonlinear diode simulation technique allows for a significantly smaller

29 calculation workload than the nonlinear diode as modeled according to the Chua or

30 Kennedy circuits. This leads to a more efficient digital algorithm for implementing

1 the component in a system, particularly when a multiple transmitter/receiver system

2 necessitates the calculation of the component characteristic for a very large number

3 of different components.

4 **R. GB-ONLY RECEIVER**

5 According to certain embodiments, the inventive principles can be employed

6 without using a strictly "non-linear" circuit element. For example, the receiver non-

7 linear diode can be replaced with two functions that represent only the Gb slopes

8 5200 and 5210 as shown in FIG. 40. This receiver type will be denoted a Gb-only

9 receiver. One possible implementation for such a receiver is shown in FIG. 41. The

10 slopes are represented by two equations as follows:

11 Upper Scroll: $I = Gb*V + Ga*Vbp - Gb*Vbp$

12 Lower Scroll: $I = Gb*V - Ga*Vbp + Gb*Vbp$

13 Where the voltage at the zero current intercept is:

14 $V(upper) = +VBP(Ga/Gb) - VBP = 5240$

15 $V(lower) = -VBP(Ga/Gb) + VBP = 5250$

16 In the Gb-only receiver the Ga term is in the Gb slope equation as shown in

17 the following equations:

18 $Gb = (R2-R1)/(R1*R2) = Ga - (1/R2)$ (see FIG. 6B)

19 where R2 is elements 207 and/or 231 and R1 is elements 204 and 5360 of FIG. 41.

20 Only the Ga slope created by the nonlinear element is taken out of the receiver

21 element.

22 In the receiver, this embodiment eliminates the breakpoints 5220 and 5230

23 (see FIG. 40). This is important because noise when summed with the signal can

24 cause the incoming signal voltage to hit the breakpoint and switch momentarily to

25 the other scrolls. This can cause bit errors in the recovered signal due to voltage

26 spikes generated in the now-removed Ga region (FIG. 40). Note the load line 5270

27 is maintained with respect to the transmitter implementation and the breakpoint

28 voltage and current points 5220 and 5230 are also the same as the nonlinear diode of

29 the transmitter.

30 This receiver embodiment allows operation in noisier channels where other

1 receivers would give bit errors when points 5220 and/or 5230 were hit due to channel
2 noise. This receiver implementation also eliminates the problem associated with the
3 signal being attenuated in the channel. In a diode receiver implementation described
4 previously, when the breakpoint voltage point is reached at either 5220 or 5230 a bit
5 error can be induced in the receiver because of the slope change and the attractor
6 transitioning to the other scroll region. With the slope Ga removed (compare FIG.
7 8), the strange attractor plus noise never causes a transition between scroll regions
8 and the incoming signal is just attenuated towards the origin 5260 (i.e., the system
9 never changes regions). If a lower scroll attractor was received then it is in the right
10 half of the V plane and can be distinguished from the upper scroll attractor, which is
11 in the –V plane.

12   In FIG. 41, a voltage difference is set up across the synchronizing resistors
13 1430 and 1450 in FIG. 41 equal to the current difference associated with the slope
14 Gb. If a fixed length chaotic vector is sent and analog to digitally converted at the
15 period of the information bits (data rate) of the transmitted signal, the system can
16 distinguish between an upper and a lower scroll attractor. The upper scroll attractor
17 will be in the negative portion of the V plane and the lower scroll attractor will be in
18 the positive V plane. This technique of linearizing the receiver can be applied to any
19 chaotic system that uses a nonlinear equation set (for example the Lorenz equation
20 set). Controlling the transmitter in such a manner that it never hits a breakpoint
21 ignores the problem of the receiver's being stimulated in the presence of noise to hit
22 the breakpoint. For optimum operations, two or more single scroll attractor pairs in
23 the same region (upper or lower) can be employed.

24   The technique discussed herein to transmit chaotic vectors (i.e., bursts of
25 chaotic signal that do not cross the breakpoint voltage) when used in conjunction
26 with a Gb-only receiver allows chaotic receivers of the type shown in, e.g., FIGs. 41,
27 42, 45A, and 45B to operate in noisy channels. These Gb receiver embodiments can
28 also be used with a transmitter that uses feedback to stabilize the chaotic transmit
29 signals to insure the chaotic signal never crosses the breakpoint voltage at the
30 transmitter (i.e., stop a single scroll attractor). In a noisy system the Gb receiver has

1   no breakpoint voltage 5220 and 5230 (FIG. 40). For an upper scroll attractor, the

2   noise must drive the system into the opposite V-plane before bit errors are generated.

3   This means that the system gains the breakpoint voltage difference in receiver

4   performance. In addition, the signal can be attenuated to nearly zero volts and still

5   be detected as a voltage difference across element 1430 and 1450 in FIG. 41, 45A,

6   and 45B.

7   One detector using an analog-to-digital (sample and hold circuit) converter

8   embodiment is shown in FIG. 42. In FIG. 42, the receiver is essentially the same as

9   the dual receiver implementation shown in FIG. 25. The change is that the nonlinear

10   diode has been replaced by dual Gb receiver only implementing elements 5340 and

11   5350 (i.e., one Gb in the upper plane and one in the lower plane of the IV

12   characteristic curve as shown in FIG. 40, where the Gb values can be different

13   depending on the strange attractor pairs transmitted). The voltage sources 5310 and

14   5330 represent the voltage crossings where the current flow is equal to zero as shown

15   in the equations above for the voltage crossings. The voltage sources are shown in

16   FIG. 40 as points 5240 and 5250. The negative resistor implementation represented

17   by elements 202, 203, 223 and 204 as well as 5360 remain the same. The resistor

18   elements 207 and 231 represent the resistance of the Gb portion of the nonlinear

19   diode. Each half of the circuit represents the upper and lower portions of the Gb

20   negative slope respectively 5340 and 5350. To detect the change in attractors, the

21   detectors previously described in this document can be used.

22   In addition, a sample and hold circuit can be used to sample the signal at a

23   multiple of the data rate to obtain a plot of the voltage at 381 compared to points

24   1470 and 287. These plot the tank circuit voltage 361 against the Gb receiver

25   voltages 1470 and 287. If one has an upper scroll, then the voltage is in the negative

26   portion of the V-plane and if it is a lower scroll attractor then it is in the +V-plane.

27   It requires several volts of noise along the V axis of FIG. 40 to cause the V1 287 or

28   1470 voltages to change signs. This is a very noise immune receiver implementation.

29   This detector circuit will also detect two single scroll attractors, even if they are both

30   in the same region (i.e., upper or lower scroll region).

1   FIG. 42 shows the sample and hold circuit detector (analog-to-digital

2   converter) connected at points 1470, 287, and 281. A matrix of the voltages 5360,

3   5370, and 5380 can be generated as shown to distinguish between the two attractor

4   pairs. Another detection technique is to sample at points 1470 and 287 and correlate

5   to a stored signal to make a detection decision. Since chaotic attractors can produce

6   uncorrelated samples, the same receiver circuit can be used to detect two or more

7   uncorrelated signals. These signals can then be decoded using a correlation circuit

8   attached to points 381, 287, and 1470. The tank circuit 361 tends to remove noise

9   and provide an excellent detection technique. It gives a positive means to detect the

10   strange attractor vector pairs in noisy channels. By using a Gb only receiver the

11   voltage at point 381 and 287 and 1470 never hit a breakpoint even in the presence of

12   high noise levels in the channel. This means the matrix in FIG. 42 can provide an

13   accurate means of detection using single or multiple samples of the signals until the

14   noise drives the signal into the opposite V-plane. This provides a noise improvement

15   over other systems.

16   In the Gb receiver implementation, the Gb receiver allows a wider dynamic

17   range in the received signal level. The Gb receiver embodiment allows operation in

18   higher noise levels by a factor of the breakpoint voltage since the voltage never

19   changes regions unless the zero voltage line 5260 is crossed. The Gb receiver

20   embodiment also simplifies the receiver design since no diodes are required in the

21   receiver. It also allows uncorrelated chaotic signal vectors to be received by the same

22   detector circuit and then separated by a correlator circuit.

23   The Gb-only receiver technique can be applied to any chaotic nonlinear

24   system that has a nonlinear element or elements. The Gb receiver can detect double

25   scroll or two single scroll attractors, even if they are both in the same region (i.e.,

26   upper or lower scroll region), and can provide an accurate means of detecting chaotic

27   signals using single or multiple sample and holds of the signal in noisy channels.

28   This embodiment allows the signal to be attenuated in a channel by more than an

29   order of magnitude and still be detected by the Gb only receiver in a noiseless

30   channel. In a conventional Chua's receiver using a nonlinear diode implementation

1   the signal can be attenuated by no more than approximately 1-2 percent before the

2   breakpoint is crossed and bit errors result in a noiseless channel.

3   In a noisy channel the Gb only receiver allows the signal to be attenuated by

4   50 percent or more and still detect the signal. The attenuation of the signal does not

5   affect noise performance. The energy per bit to the noise level is the only

6   determinant when using a Gb receiver. In some receivers, the noise floor is set by

7   the breakpoint voltage based on the signal attenuation in the channel. In the Gb only

8   receiver, the transmitted chaotic signal can be attenuated to nearly zero volts and still

9   be detected as a voltage difference across element 1430 1450 in FIG. 41, 45A, and

10   45B. All filtering techniques in this embodiment apply also to the Gb-only receiver.

11   Turning to the dual Gb-only receiver embodiment shown in FIG. 45A,

12   negative resistor elements 5340 and 5350 of FIG. 41 are replaced with negative

13   resistors 5345 and 5355, and voltage sources 5310 and 5330 are placed in series with

14   resistors 5366 and 5365 respectively. The other aspects of the circuit as similar to

15   that shown in FIG. 41. Resistors 5366 and 5365 are in series with operational

16   amplifier 223 that converts resistors 5366 and 5365 into negative resistors by

17   changing the direction of current flow. Resistors 5366 and 5365 are the parallel

18   combination of resistors 5366 and 204 and resistors 5365 and 5360 respectively of

19   FIG. 41. The resistors 5366 and 5365 are determined in the same way as in the other

20   embodiments described herein.

21   FIG. 45B shows yet another embodiment in which resistors 231 and 207 are

22   inserted in parallel with negative resistor elements 5366 and 5365 respectively. This

23   makes up the resistive elements 5346 and 5356 respectively. In this case, the parallel

24   combinations of the resistors are determined in the same way as FIG. 41. The

25   difference between FIG. 41 and FIG. 45A in this embodiment is that voltage sources

26   5310 and 5330 are in series with the negative resistor chain made up of elements 223,

27   5366, 5310 and 223, 5365, 5366 respectively. In this embodiment, elements 5345

28   and 5355 in FIG. 45A are replaced by elements 5346 and 5356 in FIG. 45B. In FIGs.

29   41, 45A and FIG. 45B one starts by designing a standard Chau-based receiver as

30   discussed above and translates the Gb slope into the negative Gb slope receiver to

1   obtain noise immunity. These circuits provide 6-9 dB better performance than a

2   standard nonlinear receiver with slope Ga present.

3   S. **GB-ONLY TRANSMITTER**

4   According to certain embodiments of the invention, chaotic transmitting

5   circuits exhibiting a linear (instead of nonlinear) current-voltage characteristic can

6   be employed, wherein the slope of the curve is perturbed to transmit information.

7   FIG. 43 shows a transmitter similar to FIG 6C except the non-linear diode 680 has

8   been replaced with the Gb slope element 5460. This transmitter produces single

9   scroll attractors in either the lower quadrant of the characteristic curve of FIG. 40

10 (curve 5200) or the upper quadrant (curve 5200) depending on the value of the

11 voltage source 5420. Voltage source 5420 represents point 5240 or 5250 in FIG. 40.

12 The single scroll strange attractor orbits around the equilibrium point where load line

13 5270 crosses Gb line 5200 or 5210 in FIG. 40.

14   The resistive element 5410 has the same value as resistor 686 FIG. 6C. The

15 resistor 204 is the same as resistor 204 in FIG. 6C. All other parts of the circuit are

16 similar to those shown in FIG 6C. The Gb slope can be modulated by placing a

17 switching element 235c and a resistor 206 across resistor 5410 and/or 206 as shown

18 in FIG. 43. The voltage source 5420 is varied as the slope is changed based on the

19 same equations as the equations for the Gb receiver discussed with reference to FIG.

20 41. The resistor 206 will be a different value depending on whether it modulates

21 resistor 5410 or 206. The input can be modulated by a digital data stream at point

22 236 (see FIG. 43).

23   FIG. 44 shows a dual Gb transmitter. It is similar to the dual transmitter of

24 FIG. 24 except the nonlinear diode subsections have been replaced by Gb-only

25 negative resistance sections 5500 and 5510. The upper Gb curve 5200 is

26 implemented in subsection 5500 and the lower Gb curve 5210 is implemented in

27 subsection 5510 as shown in FIG. 40. The voltage sources 5520 and 5530 represent

28 points 5240 and 5250 on FIG. 40 respectively. The switching subsection 1265,

29 summing amplifier subsection 1295, filter 1310, and attenuator subsection 1360

30 perform the same function as in FIG. 24 of switching between the upper and lower

1   Gb subsections 5540 and 5550 and buffering the signals 1350 before they are

2   injected into a baseband channel of a transceiver. The transmitter filtering methods

3   and circuit elements described previously are applicable to the Gb-only transmitter

4   embodiments and are incorporated therein.

5   **T. POSITIVE SLOPE TRANSMITTERS AND RECEIVERS**

6   According to other embodiments of the present invention, positive-slope

7   circuit elements are substituted for various negative-slope elements described

8   previously. Some of these embodiments are illustrated in FIGS. 46 through 52.

9   FIG. 46 shows a positive slope (Gb+) current-voltage characteristic curve for

10   resistive elements 5600 and 5610 of FIGs. 47 and 48 (the characteristic curves are

11   lines 5283 and 5290). Both lines have a positive slope as opposed to a negative Gb

12   slope (compare with FIG. 40).

13   To design positive slope curves as shown in FIG. 46, the techniques discussed

14   for the chaotic systems discussed herein can be applied to determine the maximum

15   and minimum circuit components as shown in FIGS. 19A through 19F. The

16   difference is that the sign of Gb is changed from a negative sign to a positive sign.

17   This shifts the –Gb slopes 5200 and 5210 of FIG. 40 by ninety degrees. The -Gb

18   slopes are rotated about the break points 5220 and 5230. The intersection of the +Gb

19   slope with the load line 5270 defines asymptotically stable points to a specific

20   voltage level applied to the input to the receiver of FIG. 47.

21   Any incoming voltage in the receiver that is not matched to these voltage

22   points generates a voltage difference across resistors 1450 and 1430 in the

23   embodiments of FIG. 47. In effect, this represents a high "Q" circuit to matched

24   voltage levels established by the intersection of lines 5283 and 5290 with the load

25   line 5270 at the input to the Gb+ receiver of FIG. 47. The voltage cross points with

26   the load line can be changed by moving the breakpoint voltage points 5220 and 5230

27   in FIG. 46 on the voltage axis thus changing the voltage sources +V 5240 and -V

28   5250 as shown in FIG. 46. This can be done by changing voltage sources 5310 and

29   5330 of FIG. 47.

30

1   Voltage sources 5310 and 5330 in the receiver of FIG. 47 are changed to

2   match these voltage crossings in accordance with the equations for FIG. 41 above,

3   except the sign of Gb remains positive to create the positive slope Gb+. M-ary

4   voltage levels can be created where elements 1370 and 601 are reproduced 2^N times

5   as in the M-ary waveform discussion above. In this case, the discriminating element

6   is the equilibrium point voltage 5281 and 5282 formed by the high "Q" circuit

7   consisting of tank circuit 361 and resistive elements 5600 and 5610. This receiver

8   system, when coupled with the transmitters of FIG. 49 and FIG. 50 and the detector

9   circuits discussed above, operates within 0.25 dB of the Eb/No curve for coherent

10   optimum BPSK. However, it has characteristics that distinguish it from BPSK.

11   First, the receiver does not need a phase circuit. Second, the receiver still works

12   through a filter even with 30% of a bit period time delay. Third, M-ary level

13   operation is possible with the break point voltages adjusted. This system provides

14   a self-synchronizing system while BPSK is not self-synchronizing.

15   FIG. 48 shows a Gb+ implementation of the dual receiver of FIG. 42 using

16   a positive Gb slope characteristic curve implementation as shown in FIG. 46. The

17   system can be implemented in a similar manner to that of FIG. 42. The analog-to-

18   digital converters perform the same function as they do in FIG. 42 above. This

19   implementation allows a digital system to sample and process the signal in a

20   computer.

21   FIG. 49 shows a positive Gb slope transmitter. This embodiment of a Gb+

22   transmitter can be modulated by switching resistor 5650 using a switch 235c and a

23   resistor in parallel with resistor 5650. This changes the equilibrium point voltages

24   5281 and 5282 crossings on the load line of FIG. 46 without changing the breakpoint

25   position points 5220 and 5230. The Gb+ slope rotates about the old breakpoint.

26   Another way to modulate the transmitter is to change the voltage source 5675. This

27   source moves the zero current crossing of the characteristic curve without changing

28   the slope of the Gb+ element as shown in FIG. 51 (i.e., the breakpoint voltage is

29   moved). Also, the new Gb+ line is parallel to the old Gb+ line. At the same time,

30   it changes the Gb+ characteristic curve crossing point of the load line in FIG. 46.

1 This sets new equilibrium points. This allows M-ary equilibrium points to be set as

2 discussed below.

3 FIG. 50 illustrates the use of a digital-to-analog converter to implement a

4 positive Gb+ slope transmitter. In this embodiment, the Gb+ voltage levels are

5 selected along the load line of FIG. 51 in accordance with the voltage levels that can

6 be generated by the digital to analog converter (D/A) converter. This yields 2^N

7 voltage levels where N is the D/A converters' binary level capability. There are

8 currently D/A converters capable of 2 to 16 levels. This embodiment allows levels

9 to be set to less than 0.01 volts or better.

10 The impulse response can also be applied through the D/A converter to

11 exactly mimic a Gb+ transmitter. This D/A implementation confirms that the system

12 looks like baseband BPSK with a direct current offset at the transmitter. It is at the

13 receiver that the new detector characteristics are manifest. The receiver system

14 becomes a high "Q" matched filter and synchronizes on the equilibrium points in a

15 binary or M-ary implementation. The ability to operate with 2^N levels shows that this

16 system is not just a BPSK receiver as discussed above. It has significant new

17 characteristics (i.e., immunity to phase distortion and signal delay).

18 The receiver of FIG. 47 is designed to have the same voltage crossings where

19 there are 2^N receivers for each tank circuit. There can also be multiple tank circuits

20 with multiple loadlines 5270 to change the equilibrium points. The circuit elements

21 can be selected on the basis of various chaotic circuit implementations discussed

22 herein with respect to the first and second-generation systems. The ability to define

23 the equilibrium points allows large numbers of high "Q" circuit elements to be

24 generated. The M-ary word sets can be based on equilibrium voltage levels instead

25 of waveforms as discussed herein. The reaction of the receivers FIGs. 47 and 48 to

26 a change in the incoming voltage level allows the energy per bit to noise per hertz of

27 bandwidth for these non-coherent receivers to come within 0.25 to 0.5 dB of the ideal

28 coherent binary phase shift keying systems. This results in a significant

29 improvement in performance. This modulation technique will be referred to as

30 "chaotic impulse response modulation coding" and the M-ary coding will be referred

1 to as "SAIC amplitude impulse coding." The receiver acts like an impulse response
2 circuit that quickly achieves the equilibrium level of the incoming waveform. In
3 FIG. 50 the computing element 5700 generates an information signal and loads the
4 D/A converter with an M-ary word 5710. The D/A converter 5720 then converts the
5 M-ary word 5710 to an analog output 5730 where there are 2^N possible amplitude
6 levels 5730 based on the capability of the D/A converter 5720. The output is
7 interfaced to the channel by the channel interface circuit 5740. Element 5740 is the
8 channel interface circuits discussed in FIGS. 31 and 32. The output goes to the
9 channels designated by 5750.

10 FIG. 51 shows one implementation of an M-ary SAIC coded signal. There are
11 1 to 2^N level points 5271 through 5273. For example, the Gb+ slopes 5250 and 5283
12 set equilibrium points 5271 and 5272. Equilibrium point 5272 generates a voltage
13 5267 on the V-axis. This voltage is the voltage that is transmitted over the radio
14 channel and represents the SAIC coding of the modulating signal that changes the
15 voltage source 5675 in FIG 49 or is produced by the D/A circuit 5720 in FIG. 50.
16 The load line can also be varied to change the voltage crossings with the Gb+ curves.

17 FIG. 52 shows an example of a Gb+ slope being varied to change the
18 equilibrium point; in this case, resistor 5650 FIG. 49 is varied within the chaotic
19 stable range discussed above while breakpoint 5210 remains at the same position. As
20 the Gb+ slope is varied, the equilibrium point is moved from 5272 to 5277 and the
21 V-axis crossing is moved from 5285 to 5286. This M-ary method can be
22 implemented by increasing the number of elements 5610 and 5600 in FIGs. 47 and
23 48 to 2^N, one for each level to be detected. The shifting of the Gb+ characteristic
24 curve and the load line is done by moving the breakpoint voltage across the –Gb
25 slope line 5200 and 5210 of FIG. 52.

26 **CONCLUSION**

27 Thus has been described various systems, methods, and apparati for
28 modulating and demodulating chaotic signals. It is apparent that many of the
29 embodiments can be implemented using digital signal processing techniques rather
30 than analog circuits or discrete elements. Consequently, the claims should be

1   interpreted to encompass such circuits and elements without limitation. No claim

2   limitation appearing in the claims of this application should be interpreted to be in

3   "means plus function" format unless it explicitly recites "means for" performing a

4   specified function.